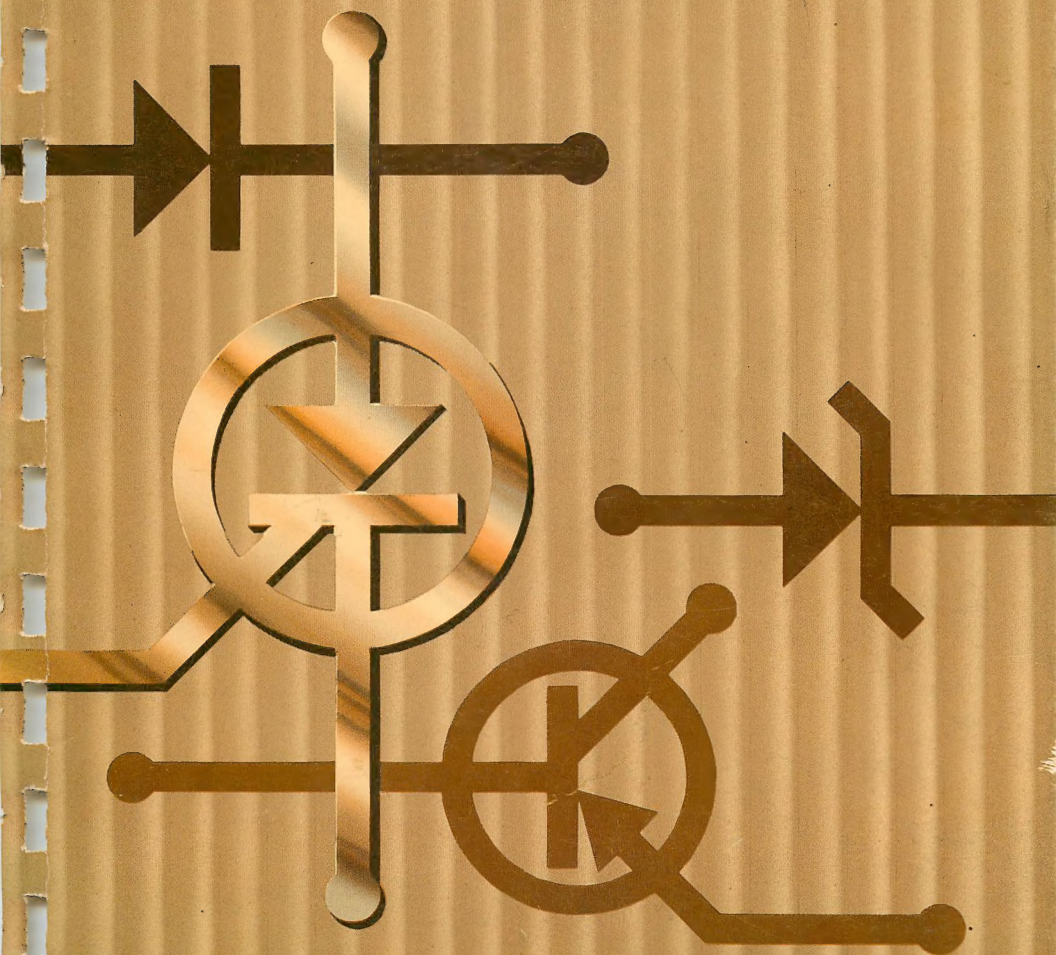


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Motorola Semiconductor

CIRCUITS MANUAL





MOTOROLA
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CIRCUITS MANUAL

- * New Products
- * Design Limitations
- * Power Control
- * Inverters
- * Regulators
- * Ignition
- * RF Transmitters

FOREWORD

This manual discusses some of the many practical circuits using power semiconductors. The circuits are intended to illustrate design principles and limitations of solid state devices. A broad frequency spectrum is covered -- dc to 240 Mc. The devices include p-n junctions such as low current rectifiers, zener diodes and the new high current multicell rectifiers; p-n-p devices including large junction germanium power transistors suitable for motor control as well as the latest silicon-epitaxial-annular-passivated RF amplifiers; PNP silicon controlled rectifiers and gate controlled switches such as the steel can low-cost SCR; and completing the circle back to a p-n junction -- the varactor diode.

The object of this manual is to provide device and circuit information which is supplemental to data sheets, application notes, handbooks and other existing technical knowledge. We sincerely hope you find this reference manual a useful guide to practical circuitry.

The information and circuit diagrams in this manual have been carefully checked and are believed to be entirely reliable but no responsibility is assumed for inaccuracies. Moreover, complete information is not necessarily given for construction purposes. Neither the disclosure of any information contained herein nor the sale of semiconductor devices conveys any license under the patent rights of Motorola, Inc. or others.

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SECTION 1

The MCR1304 Controlled Rectifier

INTRODUCTION

The Motorola MCR1304 series of controlled rectifiers are suitable for applications requiring blocking voltages through 400 volts and RMS currents up to 8 amperes. These devices are packaged in a space-saving, economical steel can which provides considerable versatility of mounting. These devices are also available with modified packages, i. e., MCR1305 series with a mounting stud, MCR1604 series which is a single-ended three-lead case and the MCR1605 series which is a double-ended three-lead case.

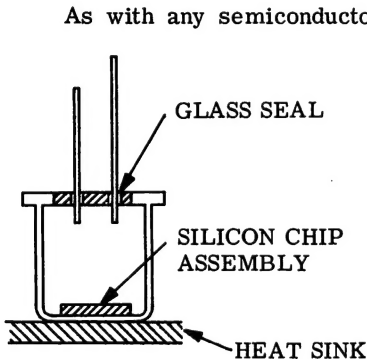


Figure 1-1 Placement of Silicon Die in Controlled Rectifier Can

die to a heat sink. Obviously, the most effective method to heat sink these devices is to mount the bottom directly to the heat sink.

As with any semiconductor device, these silicon controlled rectifiers must be operated within specified ratings for optimum and reliable performance. Semiconductor construction techniques require the use of solder as a bonding material in the controlled rectifier. To prevent damage to the processed silicon controlled rectifier die, hard solder with a sufficiently low melting point is used. Thus, in any processing of the completed controlled rectifier, the temperature must be kept below a predetermined value to prevent the solder from melting.

As shown in Figure 1-1, the silicon controlled rectifier die is mounted directly to the bottom of the case which provides the shortest thermal path from the

CONDUCTING CHARACTERISTICS

To keep within specified characteristics, the MCR1304 controlled rectifier junction, during operation, should not be allowed to exceed 100°C. The electrical limitation on this unit is 8 amperes (RMS). Imposing these two limiting values and the inherent values of thermal resistance and conducting voltage drop, the current-temperature characteristics can be determined, (Figure 1-2). Applying these characteristics to any operating condition is basically easy. For an operating current, (below the allowable maximum, determined by both magnitude and conduction angle) the temperature measured on the case of the controlled rectifier must be below the value determined from the current-temperature characteristics. Usually, either the dc condition at 180°C conduction angle state will be the worst-case operating condition because the average current drops more rapidly than the device limits with decreasing conduction angles.

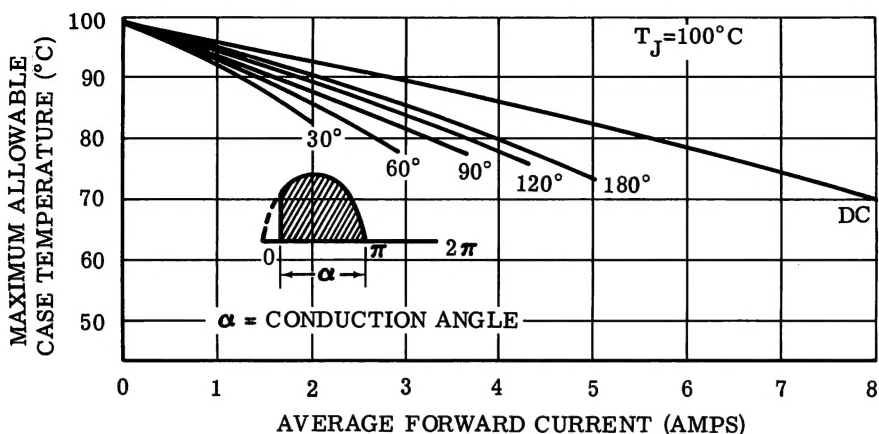


Figure 1-2 MCR1304 Current-Temperature Characteristics

For example, if it is desired to conduct 3 amperes (dc) through this series of controlled rectifiers, an adequate heat sink to maintain the case temperature below 90°C must be provided.

Also of interest to the design engineer is the power-current characteristics shown in Figure 1-3. These characteristics must be used to determine, analytically, the heat sinking requirements for the controlled rectifier circuit. From Figure 1-3, using the previous current value, we determine that the heat produced by 3.7 watts must be removed from the MCR1304 to insure proper operation. Experimentally, the design must be tested to insure proper operation, i.e. the case temperature should remain below that determined from Figure 1-2, (90°C in the example).

HEAT SINKING REQUIREMENTS

The versatility of the MCR1304 steel-can package allows it to be mounted in a wide variety of ways to meet individual requirements. Depending upon the thermal resistance value between the SCR case and a heat sink, any mounting method which satisfies the current derating curves may be used. Possible mounting media include: solder, epoxy cements; clips (fuse, resistor, transistor, special); clamps; commercial or special dissipators, retainers, coolers, and radiators.

A figure of merit which may be easily applied to any heat sinking arrangement is thermal resistance which is defined as the resistance to heat flow from a point at one temperature (T_1) to a point at a different temperature (T_2). The heat is produced by the power dissipated in the semiconductor (power = BTU/min). Mathematically, thermal Resistance = $(T_2 - T_1) / \text{Power}$. The lower the thermal resistance the easier it will be to conduct the heat away, and the better will be the heat sink.

As previously discussed, the heat sinking requirements must be determined from the power-temperature characteristics of the controlled rectifier. Combining the MCR1304 characteristics and the definition of thermal resistance, we can determine heat sink requirement data, (Figure 1-4), in terms of heat sink thermal resistance.

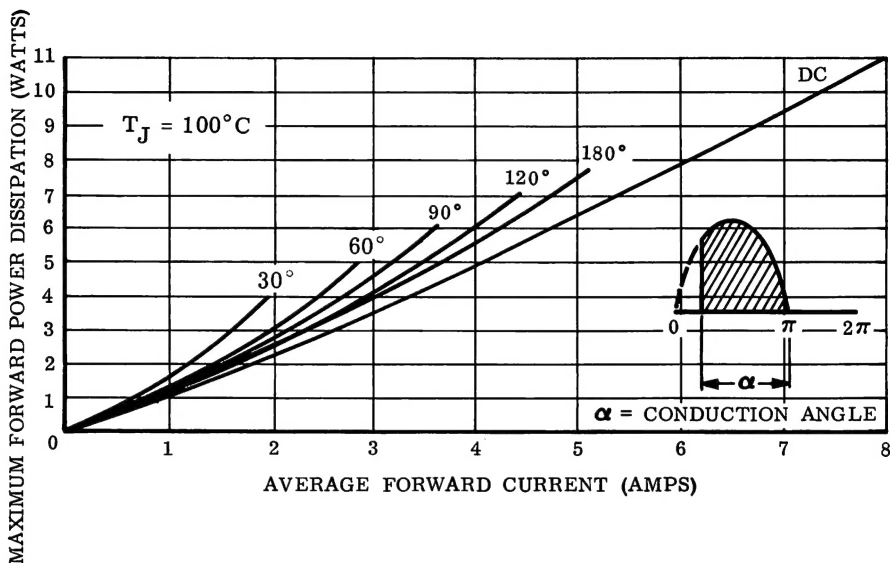


Figure 1-3 MCR1304 Power-Current Characteristics

SOLDERING THE MCR1304

One of the easiest and most efficient methods of heat sinking this controlled rectifier is to solder it directly to the heat sink. The following outline has been prepared as a step-by-step guide to soldering:

1. Materials Required:

a. Solder:

To prevent damage to the device, a relatively low temperature solder with a melting point between 175°C and 225°C must be used. The common lead-tin alloys are recommended since they are inexpensive, readily available, and they have a wide range of melting points in the temperature range required. Any lead-tin solder of less than 70 per cent lead content may be used. The 60/40 lead-tin alloy (melting point of 188°C) is the most practical in that it is generally a common stock item.

b. Flux:

Depending on the soldering method used, a flux may or may not be used. However, when used, a non-corrosive resin flux is recommended.

2. Soldering Methods:

Depending on specific needs, soldering can be effected by using either hot plate, oven, or belt feed furnace. In all cases, temperature must be controlled.

a. Hot Plate:

The hot plate is probably the most effective and flexible method of soldering. The recommended method is:

1. Set surface temperature of hot plate to a maximum of 225° C.
2. Place heat sinks on hot plate for approximately five minutes.
3. Place a 1/8" - 1/4" piece of solder on area of heat sink to be soldered.
4. As soon as solder becomes liquid, place device on this area applying slight pressure and rotating the device slightly to assure a good contact.
5. Remove the heat sinks from the heat source and cool.
6. Flux may be used if required. However, suitable wetting is achieved mechanically when the device is rotated in liquid solder.

b. Oven:

For an inert atmosphere such as N₂, dry air, etc., a flux is recommended. If H₂N₂ is available and used, flux should not be required. Again, temperature must be controlled.

c. Belt Feed Furnace:

The same procedures apply here as in the oven with the exception that possibly jiggling may be required to hold the device and the heat sink in the proper position.

3. Heat Sink Material:

- a. Since copper and most of its alloys present no problem in soldering, they probably are the most favorable heat sink material.

- b. Stainless steel is difficult to solder. However, by using a strong acid flux filled preform, satisfactory soldering can be achieved.
- c. For materials which do not solder too easily, such as aluminum, applying a tin coat to the material will usually simplify soldering.

MCR1305 - THE STUD VERSION OF THE MCR1304

For both prototype work and for users whose needs dictate a screw down unit, the steel can has been soldered to a 7/16" hex stud.

Care should be taken if this unit is to be subjected to high temperatures during circuit construction. To prevent unsoldering the case from the stud, the case of the controlled rectifier should not be allowed to exceed 150°C.

The heat conducting properties of this unit do not differ greatly for the original steel can unit. And if case temperature is measured on the side of the steel can the original power rating characteristics may be applied.

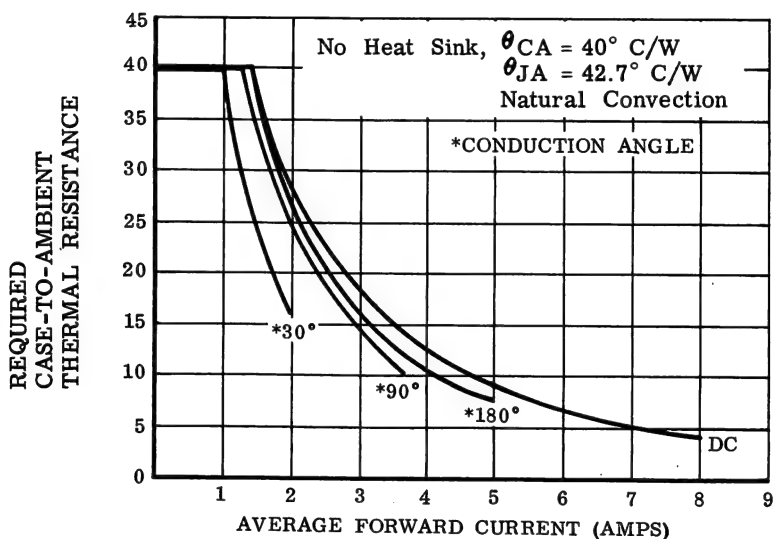


Figure 1-4 Heat Sink Thermal Resistance Requirement for $T_A = 25^\circ\text{C}$

USING THE MCR1304 WITHOUT A HEAT SINK

If desired, the MCR1304 may be used with no heat sink attached. Naturally, the current capability of the device is reduced. The thermal resistance, from the device to the air, is 40.0°C/W. Thus, to maintain the junction temperature below 100°C the device case temperature must still be kept below the temperature specified on the MCR1304 power rating characteristics. As shown in Figure 1-4, the maximum current capability is 1.5 amperes at an ambient temperature of about 25°C.

Commercially Available Heat Sinks

A survey of commercial heat sink manufacturers was conducted and a number of adequate heat sinks were found and evaluated.

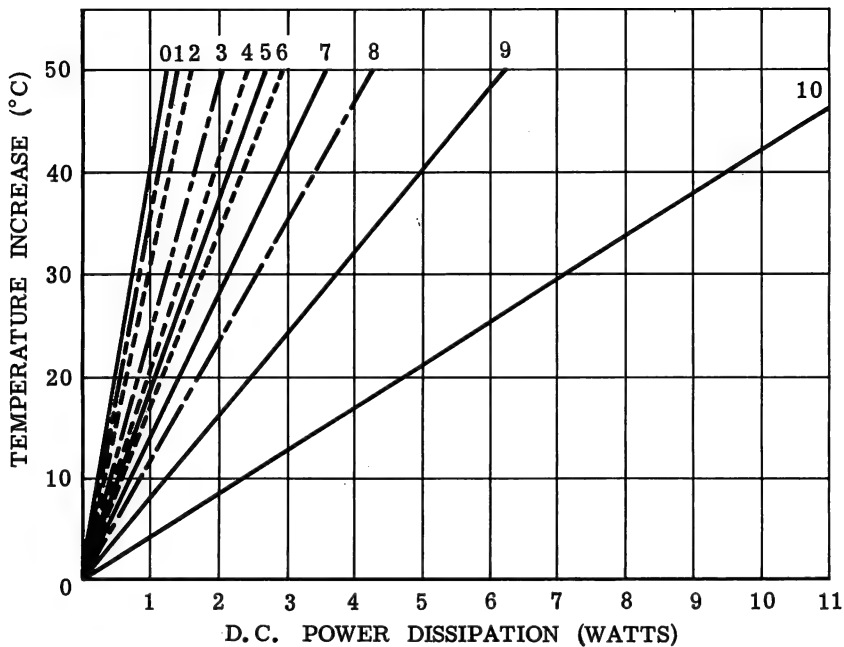


Figure 1-5 Commercial Heat Sink Evaluation Data

Figure 1-5 shows the relative merit of the evaluated heat sinks. The numbers on the curve correspond to the "operating curve" numbers listed in the heat sink table, (Table 1-6). Curve "0" represents the SCR operating in free air with no heat sink attached. Curve "10" represents the SCR soldered to a large copper plate (5" x 5" x 1/8").

For each of the curves we can obtain an effective case-to-ambient thermal resistance, by calculating the slope of the curve. Table 1-7 gives the thermal resistance for each curve.

TABLE 1-6

<u>Manufacturer & Type No.</u>	<u>Operating Curve</u>
Augat, Inc. (#9017-1P1V)	#2
Daedalus (#CS-3)	#1
Atlee Corp. (#106-707-2)	#1
with #100-236-6 clip	#2
Birtcher (#3AL680)	#1
Wakefield Eng. (#NF209)	#3
Thermalloy (#2215)	#2
Relco Prod. (Series L)	#3
Staver (V2-1 (modified)	#6
Staver (V1-3)	#7
U. S. Heat Sink (#a1041)	#4
Astro Dynamics (#2703)	#6
Astro Dynamics (#2701)	#3
Astro Dynamics (#2702)	#3
Astro Dynamics (#2801)	#3
Astro Dynamics (#2802)	#5
Staver (F3-1)	#3
IERC #LP5C1B	#3
IERC #TXS on 2 x 2 plate	#7
IERC #HP3 (with TXS)	#9

TABLE 1-7

Effective Thermal Resistance

Curve #0	$\theta_{CA} = 40^{\circ}\text{C/W}$
1	37
2	31
3	24
4	20
5	19
6	17
7	14
8	12
9	8
10	4



SECTION 2

The Gate Controlled Switch

INTRODUCTION

The solid state controlled rectifier (SCR) behaves similar to the vacuum tube thyatron in that turn-off is accomplished by reversal of voltage polarity from anode to cathode. Actually, in the SCR as in other four-layer devices, turn-off is accomplished when the anode current falls below what is known as "Hold Current", hence, the voltage applied to the gate has little or no effect in turning the device "off".

The gate controlled switch (GCS) has the ability to turn off when a negative voltage of sufficient value is applied to the gate. The GCS is analogous to the type of mechanical switch which closes when pushed once and opens when pushed again. These characteristics differentiate the GCS from a power transistor, which requires a constant signal at its base to keep it "on" as well as a constant signal of opposite polarity applied to the base to keep it in the "off" condition. The GCS can also be operated as an SCR in that it can be turned off by anode-cathode voltage polarity reversal.

The turn-on and turn-off mechanism can be considered the same from the following expression for anode current:

$$I_A = \frac{I_S \pm \alpha_1 M I_G}{1 - M(\alpha_1 + \alpha_2)}$$

where I_A = anode current

I_G = gate current (+ or -)

I_S = forward leakage current

α_1 & α_2 = the respective equivalent current gains of the two transistor analogy

M = voltage multiplication factor (1 to ∞)

Obviously if the term $M(\alpha_1 + \alpha_2)$ equals the quantity "one" the anode current will suddenly avalanche and be limited only by external resistance. This can occur by the increase of "M" due to voltage or by increasing $(\alpha_1 + \alpha_2)$. At low voltages $(\alpha_1 + \alpha_2)$ is very close to unity and not much increase is necessary to initiate turn-on. Since current-gain can be increased by injecting gate current, turn-on can be easily accomplished.

However, turn-off is more difficult. The anode current equation still holds true and if gain can be decreased to a low enough value a regenerative turn-off will occur. If enough negative gate current can be applied, the gain can be decreased. Unfortunately, the amount of gate current may be quite large since the gains at high anode current approach high levels and a special method must be used to allow practical gate turn-off. Any SCR may possibly be turned-off at the gate but this is rare. The GCS is deliberately constructed to allow gate turn-off. Device engineering indicates that α_1 should be at least .95 or more and α_2 should be about .02 to achieve a turn-off gain greater than 5.

Another way of looking at this phenomenon is that increasing negative gate current increases hold current until the device turns off.

The Motorola MGCS821, MGCS-924 and MGCS-925 are gate controlled switches which have been designed to have a turn-off gain of 10 at 5 amperes.

GATE CHARACTERISTICS

The GCS static forward characteristics are the same as an SCR. The important difference is evident in the V-I characteristics of the gate-cathode terminal pair as shown in Figure 2-1.

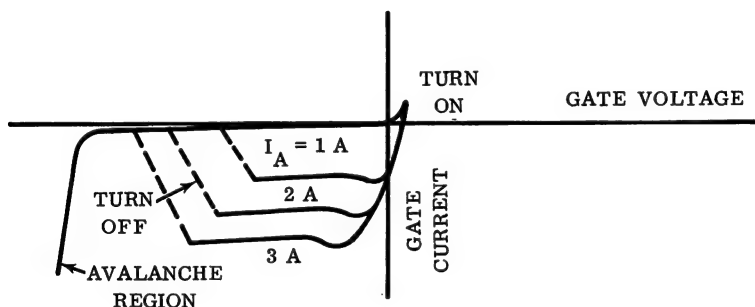


Figure 2-1 GCS Gate-Cathode Characteristics

The SCR "anode-on" gate characteristics would follow very closely the "off" characteristic of the SCR or GCS if the gate is swept with a negative voltage. However, the "anode-on" gate curve for the GCS shows a great increase in negative current and then flattens out as the applied source voltage increases in the negative direction until suddenly the curve becomes discontinuous and the operating point will jump to the off curve as dictated by the source resistance. If the source voltage is allowed to increase more the curve would trace out the avalanche region. And if the voltage is decreased with the anode off, the curve is simply the reverse characteristic of a back-biased junction.

SWITCHING TIMES

The GCS can turn-on to 5 amperes in about one micro-second. During

turn-off the device exhibits storage time, fall time and hold time as shown in Figure 2-2. Gate current and gate voltage waveshapes are also shown. The source voltage was rectangular and a 10 ohm source resistance was used.

The storage time can be considerably reduced by overdriving the gate with a low "forced off" gain. This is shown in the dotted portions of Figure 2-2. To do this requires high source voltages and when the device turns off the gate operating point will be intercepted in the avalanche portion of the V-I curve. However, if the gate drive is removed within 50 microseconds or so, there will be no damage done.

The fall time is about 2 microseconds at 5 amperes anode current and the hold time may be as high as 10 microseconds although this has not been specified as yet. Overdriving does not materially decrease the fall or hold times. The significance of hold time is evident if gate signal is removed too soon the anode will switch back on. Thus, the off gate signal has a minimum applied time equal to the total forward current turn off time (storage + fall + hold) or the device reverts back to conduction, no matter how hard the off drive may be.

There are many circuits which can be used to turn off a GCS; dc, ringing choke, capacitor discharge, rectangular pulse, etc. All of these circuits must have the following qualifications, for the MGCS821, 924 & 925:

- 1) High enough source voltage (12 volts at least to turn off 5 amperes).
- 2) Low enough source resistance to allow proper off current.
- 3) Exist in time long enough coincident with Items 1 & 2 to allow charge equilibrium (10 to 20 μ sec.)
- 4) Low enough energy content so as not to burn out the gate. (The exact value of this has not been determined.)

An example of a capacitor discharge turn-off circuit (5 amperes) is shown in Figure 2-3.

A 2 μ f capacitor is charged to 40 volts and then discharged through a 20 ohm resistor into the gate. We can see that this combination fulfills the four requirements. At the end of one time constant (40 μ sec.) there will be about 15 volts and, hence 750 mA still available for turn off. In fact, the device will be completely off before one time constant. The total energy is only 1600 micro-watt-seconds and most of this will be dissipated in the 20 ohm resistor.

Inductive and rectangular pulse circuits must be designed in a similar manner. The dc circuit must be designed with Point 4 especially in mind. The pulse methods may cause the gate to avalanche but will not have enough energy to cause burn out. However, the dc or low frequency turn off circuits should avoid the avalanche mode. This is easily accomplished by using a very low (1 ohm) source resistance from a 12 volt source. The gate current will then be just a few milliamperes after turn off and no gate damage will occur.

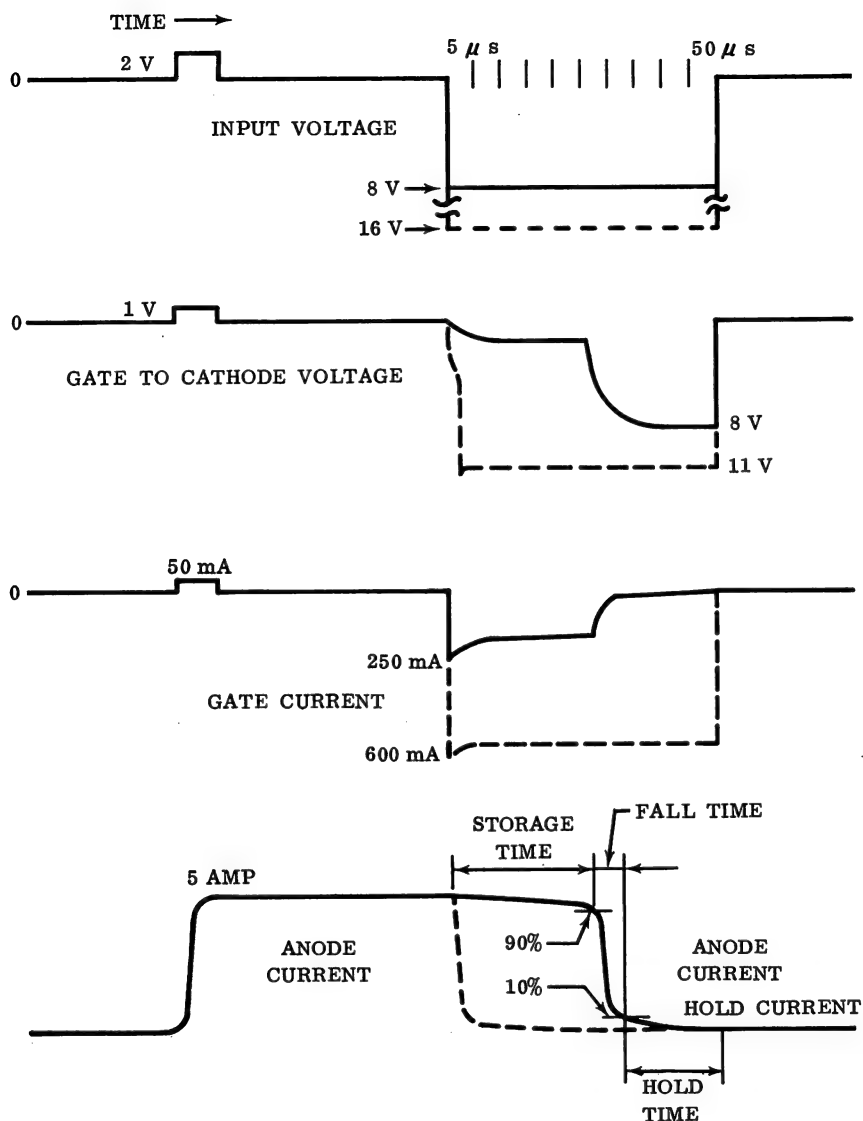


Figure 2-2 Typical GCS Switching Waveforms

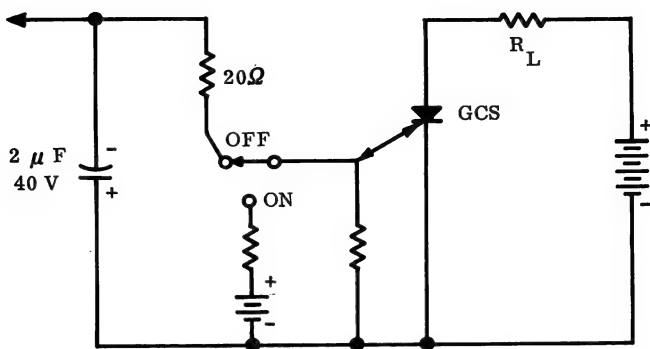


Figure 2-3 GCS Turn-off Circuit Capacitive Discharge

SAFE AREA

Although the GCS may be rated at 5 amperes and 400 volts, this does not mean that these ratings are simultaneous even during switching. Secondary breakdown may occur resulting in an anode to cathode short. A triangular safe area is defined for these devices. Section 5 includes a typical safe area.



SECTION 3

Reverse Polarity SCR'S

Just as the reverse polarity rectifier complements the forward polarity rectifier, giving the equipment designer an additional degree of freedom for simplification of mechanical and heat sink design, so does the reverse polarity SCR complement the forward polarity version. Motorola has developed the unique ability of making a reverse polarity SCR. That is, a device in which the cathode rather than the anode is connected to the case. Such devices are now available in the press-fit line, and are designated as MCR808R-1 through -6, and MCR1308R-1 through -6.

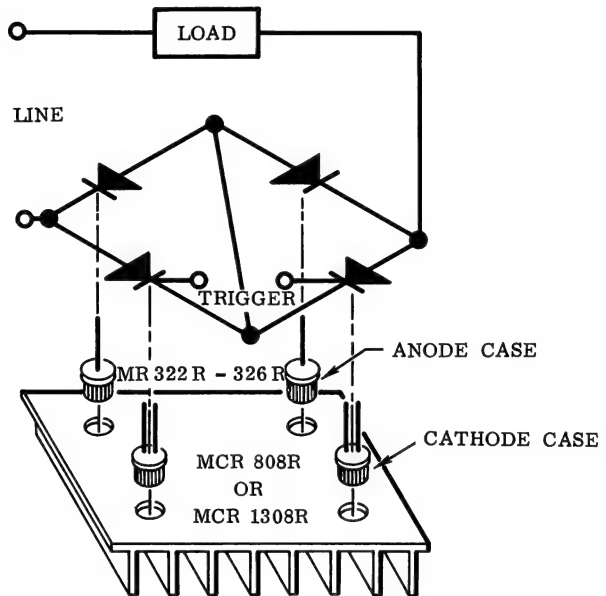


Figure 3-1 Two SCR Shorted Bridge

An example of a circuit in which considerable savings can be achieved by using reverse polarity devices is the two SCR shorted bridge shown in Figure 3-1. Note that there is a common connection between the two rectifier anodes and the two SCR cathodes. By using reverse polarity rectifiers and reverse polarity SCR's, all four devices may be pressed into the same heat sink. Of course, by interchanging the rectifiers and the SCR's, the same result could be achieved with forward polarity devices. However, with the SCR cathode's separated, separate trigger circuits or a trigger transformer would be required.

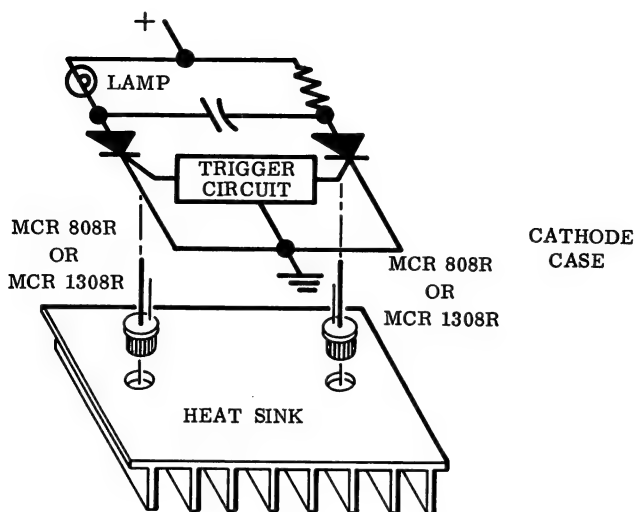


Figure 3-2 Flasher

The flasher circuit of Figure 3-2 is another example of the type of circuit which permits heat sink savings by using Motorola reverse polarity SCR's. The parallel inverter and the chopper type motor control are related to this circuit.

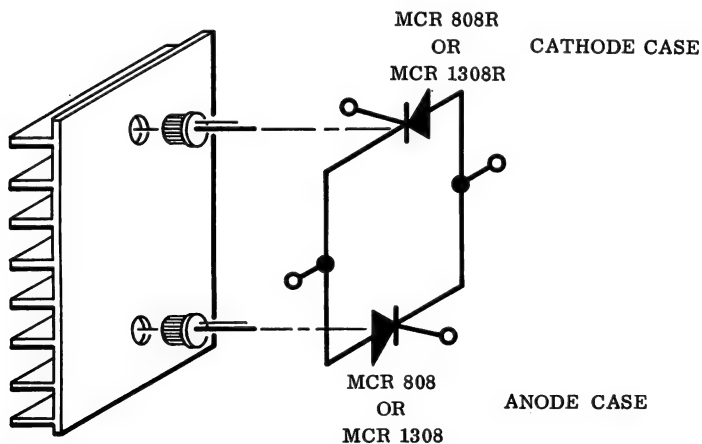


Figure 3-3 Lamp Dimmer or Motor Control

The utility of the reverse polarity SCR is not limited to those circuits requiring a common cathode connection between two SCR's. A reverse parallel circuit, such as the lamp dimmer or motor control is an example of the type of circuit in which savings can be accomplished by using a reverse polarity device in conjunction with a forward polarity device. Automotive applications, such as ignition and voltage regulator circuits can be reduced in cost because the heat sink can be part of or connected directly to, the frame of the automobile without insulation.

There are a host of additional circuits and design problems which can be simplified by the reverse polarity SCR. Keep it in mind whenever you are faced with a heat sink or grounding problem.



SECTION 4

High Current Rectifiers

INTRODUCTION

The ability of a silicon PN junction to efficiently rectify current is directly effected by the permissible current density and the series resistance of the various metals used in the package. In addition there is the thermal resistance factor which is directly related to junction area. Even if a small area junction had zero series electrical resistance and no current density limitations, there still would be a limit imposed on the forward current.

For example, assume the rectifier is perfect and has 0.8 volts drop at all current levels. Then, power dissipation would be,

$$P_D = 0.8 \times I_F.$$

And temperature rise,

$$\Delta T = P_D \theta_{JC} = .8 I_F \theta_{JC}.$$

The thermal resistance, θ_{JC} , is then the key to high current rectifiers, and large junction area is the key to low thermal resistance.

There are two basic methods to obtain large area junctions. One method uses large single rectifier junctions to carry the required current (as the current requirements are increased these single junctions must become larger and larger) and the second method is to parallel as many lower current rectifiers as needed.

In the past, the manufacture of high-current rectifiers has proceeded with the large single-junction method because of the difficulty of accomplishing equal current distribution through parallel cells without the use of external current equalizing components, and also because the cost of producing and packaging individual paralleled cells was greater than the single junction approach.

However, there are a number of problems associated with large single-junction rectifiers which make them far from being the ideal elements for high-current rectifiers. Large area rectifier junctions cannot be made without some junction imperfections, and the larger the area (for higher current capability), the greater the chance of imperfections. Moreover, even though they are referred to as large single junctions, in actuality they consist of many small junction areas intimately connected in parallel.

The voltage drop across all of these small junction areas is not entirely the same; therefore, some areas will assume more than their normal share of current. These high-current areas may ultimately result in high heat density areas which can cause thermal fatigue and deterioration. These thermally stressed areas and other imperfections can assume excess current on surges and destroy the entire rectifier.

Instead of using one large junction rectifier, Motorola employs a number of perfect (for all practical purposes) smaller junction, medium-current units connected in parallel to provide practically any desired total current rating. Motorola produces the world's highest current rectifiers by this method. Breaking up the large rectifier junction into many small rectifier junctions permits the selection of "perfect" devices which then are paralleled to form a total junction with more ideal characteristics (lower V_F and I_R) and greater current and surge capabilities.

If the parallel cell concept is so good, why hasn't it been used before now by other rectifier manufacturers in the high-current rectifier business? The fact is that although this parallel concept has been known and accepted for some time, one basic reason made it still not practical. This reason was that not until Motorola developed the large production facilities for the basic cells (more than 20 million rectifiers are being manufactured by Motorola's facilities each year) was it practical to develop and produce in volume the suitable devices for this new concept. This high production capacity brought individual cell costs down to a practical figure for such applications. Without this manufacturing capability it would not be economically practical to select cells with such closely matched electrical characteristics.

Similar individual rectifiers (press-fit package) made with these same production facilities and techniques have been chalking up a phenomenal reliability record in rugged automotive alternator applications. Of 8 million devices used by one customer from May to December 1962, only 361 have failed in the field by the middle of 1963 -- a failure percentage of only .0045 per cent.

METHOD OF MATCHING

To overcome the basic objection to the use of multiple parallel rectifier cells (i.e. attaining equal current distribution), Motorola matches the forward voltage characteristics to within 20 milli-volts at 100 amperes (see Figure 4-1). Motorola's consistent and close process controls are such that matching forward voltage characteristics at this high current level assures an even closer match at lower operating currents.

PARALLEL CONSTRUCTION

These closely matched rectifier cells are then mounted on a common copper base in a manner which intimately couples each cell thermally (see Figure 4-2). Under normal operating conditions, the thermal differences between cells is so low that any current unbalance is negligible. Temperature gradients across the common heat sink range less than 1 or

2 degrees C. With a forward voltage temperature coefficient of about -2 mV/C, it is obvious that thermal balance is maintained.

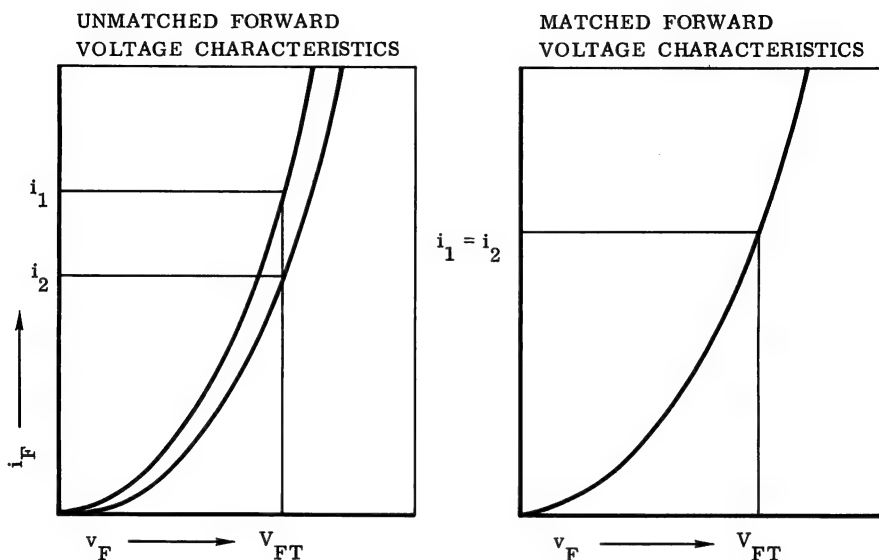


Figure 4-1 Comparison of Matched and Unmatched Cells

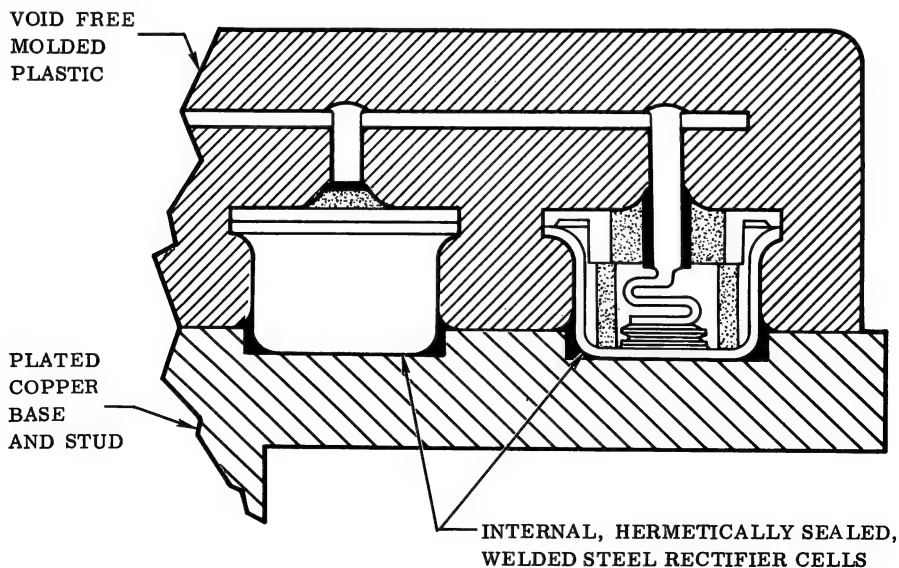


Figure 4-2 Parallel Construction

Other advantages of the parallel cell concept are: 1) the entire rectifier assembly can be factory tested prior to final assembly and any substandard cells can be replaced. This increases yields and lowers costs. 2) a number of reserve cells can be "built-in" to provide an extra current margin. If one cell happens to fail (open), the assembly can continue to function. The devices are rated such that each cell is operating at only 75 per cent of its maximum current capability.

TEST RESULTS OF MOTOROLA PARALLELED DEVICES

A university laboratory conducted tests on a 400 ampere rectifier assembly to verify that, on high-current surges, current didn't concentrate in a few of the individual cells. The following test conditions were used:

Frequency - 1 kilocycle sinewave

Surge duration - 500 microseconds

Surge - 40,000 amperes

$V_{RM(rep)}$ applied after surge

The test was repeated six times without incident, verifying that the cells conduct individually and recover fast enough to withstand large surges.

To determine exactly what happens if one cell fails, a 400 ampere rectifier was surged at rated conditions (400 amperes, 150°C case, $V_{RM(rep)}$ applied after surge) with ten consecutive 1/2 cycles of 8000 amperes. One cell began to fail after this test was performed 12 times, and within an additional three half-cycles the internal connecting lead of that cell fused open and removed the cell from the circuit. The rectifier then continued to function as a 400 ampere device. These test conditions were many times more severe than the device rating.

MOTOROLA TYPES

The present line includes 70, 160, 240 & 400 ampere devices with voltage ratings from 50 to 400. The numbers are MR1210 to MR1247. Higher current ratings are also available as specials.

An example of the permissible forward current case temperature deratings curves are shown in Section 5.

SECTION 5

Power Ratings and Safe Area Considerations of Solid State Power Devices

POWER RATINGS

All semiconductor devices have a maximum junction temperature which should not be exceeded. This rating is directly related to the maximum allowable power dissipation by the thermal resistance properties of the device and the heat sink, if used. The relation of temperature and power is usually given by what is known as a derating curve whereby the allowable power is plotted against the case temperature as shown in Figure 5-1.

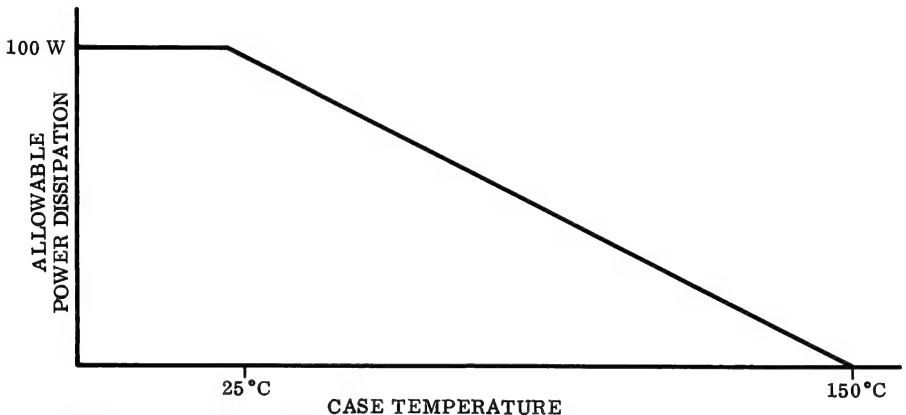


Figure 5-1 Typical Allowable Power Dissipation vs. Case Temperature

This derating curve is simple and is useful for derating average power vs. case temperature. However, depending on the device other curves are often used. For example, power rectifier data sheets invariably derate average forward current against case temperature, such as illustrated in Figure 5-2 for the 400 ampere Motorola MR1240 series.

As can be seen from Figure 5-2, more than merely average power dissipation and thermal resistance affect the allowable current. Only in the dc situation does the simple power relation hold true. Other limitations exist such as a maximum rms current. Also for operation other than dc, there are high peak currents which cause peak junction temperatures in excess of the average, thus the average allowable current must be derated accordingly. And the curves in reality are derating peak current vs. case temperature. This derating includes peak power and a concept of transient thermal resistance since the junction temperature is delayed in time with

reference to the applied power. For example, in a 180° half sine conduction of 60 cps., the peak current is π times the average value. The average power and peak power are obtained from the forward V-I curves. Then the junction temperature to a good degree of approximation is $T_j = T_c + P_{ave} \theta_{jc} + (P_{peak} - P_{ave}) \theta_{jc} \text{ transient}$.

Silicon controlled rectifiers and gate controlled switches have similar derating curves and are complicated by the fact that conduction may take place during part of a 180° sine wave, and the derating curves will include phase angle.

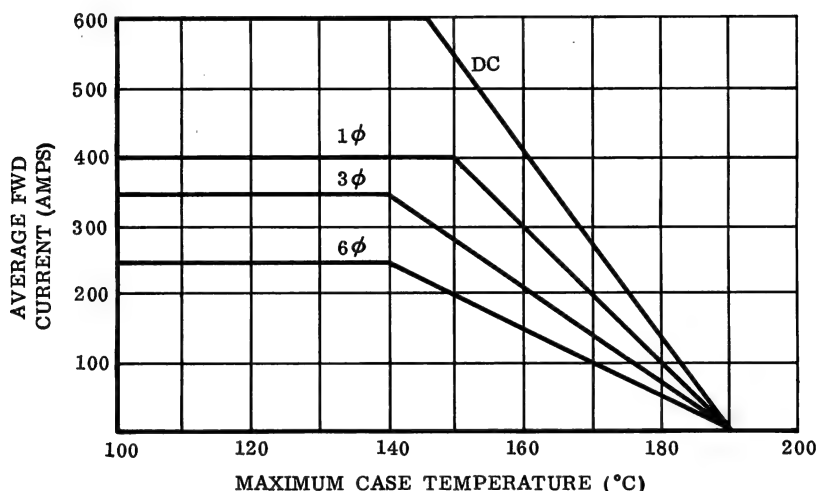
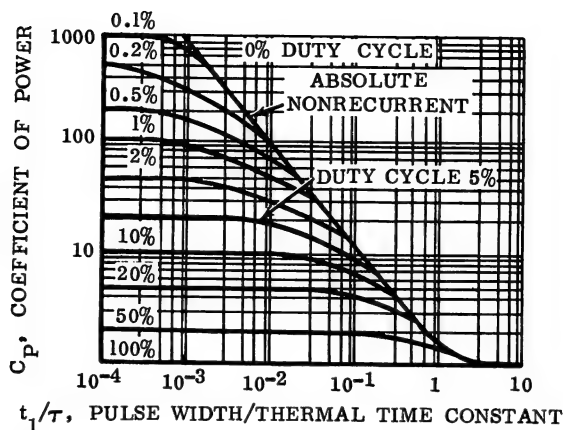


Figure 5-2 Allowable Average Forward Current Rating vs. Maximum Case Temperature

Power transistors must be derated not only as to average power vs. case temperature, but for pulse considerations as well. The thermal time constant (analogous to transient thermal resistance) for power transistors allows the construction of a pulse derating curve similar to the normalized curve of Figure 5-3. For pulse power in transistors the junction temperature is $T_j = T_c + P_{peak} \theta_{jc} / C_{pjc}$, where C_{pjc} is a coefficient obtained from Figure 5-3.

All of these curves derate power (be it pulse, average or continuous) so as to insure safe operation below the maximum junction temperature.

For rectifiers and SCR's there are certain surge current ratings based upon I^2t concepts and rms allowable which must also be considered. If these current ratings are properly used, as well as staying within the voltage ratings, a failure should never occur.



CAUTION:

In all cases the peak pulse power should stay within the Safe Operating Area.

Figure 5-3 Peak Power Derating Curve

SAFE AREA

There is another consideration for gate controlled switches and power transistors besides the power-temperature ratings. This is the "safe area" derating of allowable voltage vs. current of the output terminal pair (collector-emitter or anode-cathode). This derating is necessary since a destructive mechanism exists called "secondary breakdown". If a transistor is operated simultaneously at high currents and high voltage and for some length of time, a sudden current concentration will occur resulting in a collector to emitter short. If the GCS is switched through high dissipation regions, it will fail with an anode to cathode short. Failure analysis has led to the concept of safe areas for power transistors (See Figure 5-4) and for GCS's (See Figure 5-5).

The GCS can only be used as a switch and the safe area is simply to limit the inductive load lines during the turn-off switching time. A permissible "L" shape load line is indicated in Figure 5-5. However, a transistor can be operated at an infinite number of operating points and for an infinite number of time intervals, thus the "safe areas" include time as a parameter. Although temperature affects secondary breakdown to a certain degree (higher temperatures cause failures to occur at slightly lower power levels) the safe areas as plotted per device can be thought of as independent of temperature since a considerable safety factor is included. Both the temperature derating and the safe areas limit the permissible peak power surges. However, these are independent limitations and both must be considered in analyzing whether or not a given device is being operated in a reliable manner.

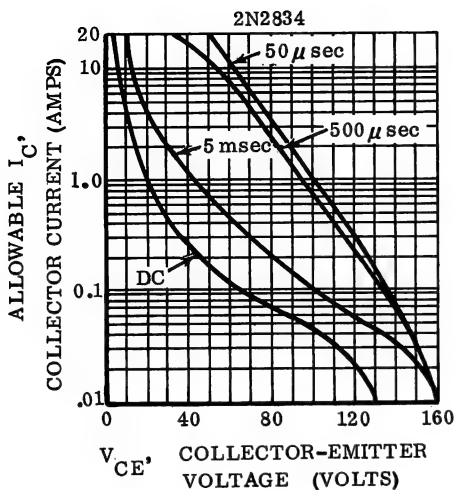


Figure 5-4

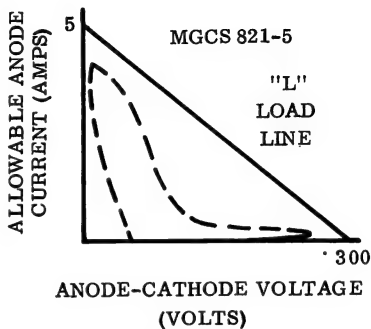


Figure 5-5

Safe Operating Areas

DC OPERATION

The safe area curves, which are time dependent, include what is called a dc safe area. Usually on alloy devices, the dc failures occur near the 25°C case temperature power dissipation area and hence, in a practical sense, under a continuous bias the device is power limited rather than safe area limited. However, with certain diffused germanium and silicon devices, the dc breakdown curve falls well below the allowable power dissipation even at higher case temperatures, based on thermal resistance and junction temperature.

An example of the dc safe limit is shown in Figure 5-6 for the 2N2832. Note that the dc limit is considerably below the 25°C case power curve, but overlaps the 100°C curve. Biasing the transistor above this dc curve may cause the device to go into secondary breakdown and a collector-to-emitter short may result, although the power dissipation is less than the allowable average power dissipation curve. The limitation then for 100°C case would follow the lowest curve and is shown as arrows following the power dissipation curve to about 2.5 amperes at 42 volts then following the dc safe area limit. At lower case temperatures, similar double limits apply.

The dc safe area only applies if the device is left on continually under some steady state bias condition. The definition of "dc" here is that "time" which is very long compared to the thermal time constant. In other words, when the device has reached thermal equilibrium. This could be many minutes on a large heat sink or a few minutes if the device is in free air.

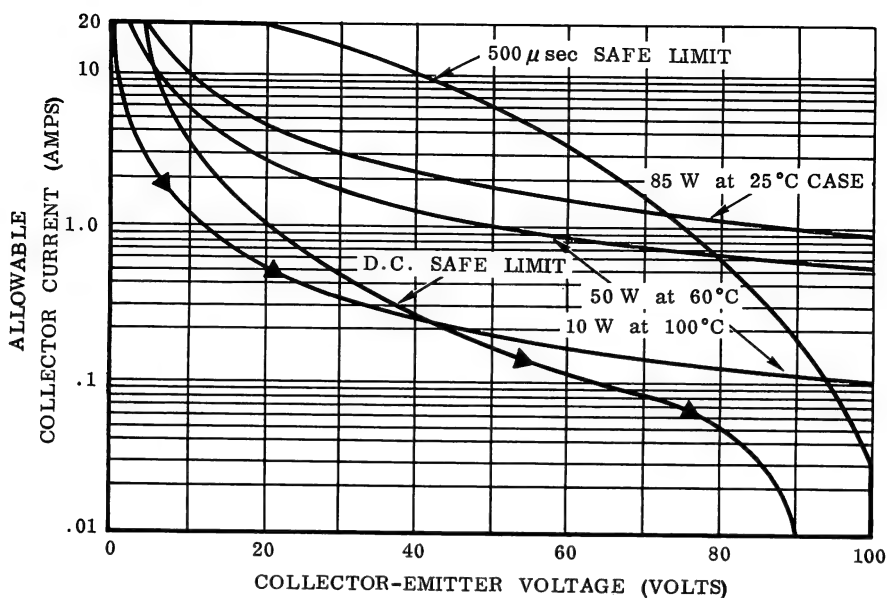


Figure 5-6 D.C. and 500 μ sec Safe Area vs. Power Curves 2N2832

An application where the dc safe area is of concern is the dc series regulator. There are conditions in a regulator where the device is safe from the dc viewpoint but not safe from an average power dissipation viewpoint. If the device is used in extremely high ambient where the case temperature is controlled to say 100°C, the device can only dissipate 10 watts of power (from Figure 5-7).

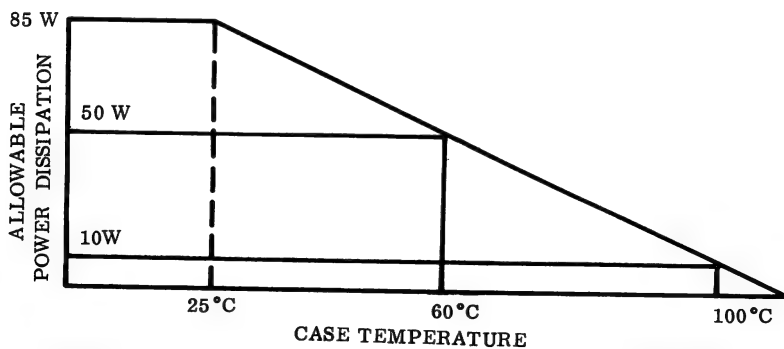


Figure 5-7 Allowable Power Dissipation vs. Case Temperature

For example, if the transistor is biased at the 10 V point, then from the power dissipation considerations, one ampere of current could be passed. In other words, the operating point is sitting at 1 ampere, 10 V continuously, which is safe from all considerations. Yet if the bias were increased to about the 2-1/2 ampere, 10 volt condition, which is permissible as to the dc safe area, but is not "power" safe, since the maximum junction temperature would be exceeded. This may or may not cause a sudden failure but it would cause some degradation in device performance.

On the other hand, operation at 80 volts would be limited by the dc safe area and only 50 mA of current is permissible or 4 watts dissipation.

PULSE OPERATION

Since the power dissipation limits are based not on continuous but rather average power, it is the average heating effect of the junction temperature that is of concern. For example, a square wave could be switched by the transistor, where the peak IC-VCE excursion is limited by the applicable safe area limit. Yet the average power must be no more than that allowable from the power dissipation curve.

Specifically, suppose the 2N2832 device is used as a pulse generator where it is biased at 50 V in the "off" state and is turned "on" with enough base drive to reach 2 amperes of collector current, still at 50 V, at a 50 per cent duty cycle. Thus, the "off" condition is 50 V at essentially zero current for 500 microseconds and is "on" at 2 A, 50 V for the succeeding 500 microseconds which is allowable from the 500 microsecond safe area as seen in Figure 5-6. This condition causes 100 W peak power to be switched but only 50 W of average power is dissipated within the transistor. From Figure 5-7, we see that an average power of 50 W is allowable if the case temperature is controlled below 60°C. An excellent heat sink would be required even at 25°C ambient, but could be a practical consideration. Thus, the peak pulse may exceed the average power curve and the dc safe area but must not exceed the specific pulse safe area.

TRANSISTOR TYPES

D. C. safe areas are not shown in most data sheets but are included in Revised Application Note #137. The 2N2832 to 2834 which are germanium devices have the dc safe areas shown on the data sheets.

SUMMARY

- 1) Allowable power dissipation is based on average power loss, maximum thermal resistance and maximum junction temperature.
- 2) Peak power can exceed the allowable average dissipation curves but cannot exceed the applicable "safe area" limit.
- 3) Collector bias at dc operation or very low frequency (times in excess of the systems thermal time constant) must be limited either by the "dc safe area" or power temperature relations, whichever is the lowest limit.

THE MYTH OF A UNIVERSAL TRANSISTOR

A study of safe area curves and power dissipation limits of both silicon and germanium power transistors reveal that a great variation exists among various products although all are encased in similar packages. The bewildering array of product leads one to yearn for a universal transistor, or, at least one with just a few compromises. Let's say at least 100 volts, 150 watts average dissipation at 25°C case, 5 microseconds safe area at 25 amps and 75 volts, dc safe area at least 120 watts throughout most of the useful range, frequency cutoff beyond 1,000 Mc, gain greater than 50 at 25 amps, and saturation resistance no greater than 40 milliohms and all of this for \$1.00.

Unfortunately, in the real world of physical science, compromises of the above specification do exist. To increase safe area certain fabrication techniques can be used which effect other transistor characteristics whether it is silicon or germanium devices; 1 - Larger junctions can be used which increase dc safe area, lowers thermal resistance which increases power rating, lowers the yield, causes higher costs, and increases output capacitance which deteriorates RF use; 2 - Higher resistivity material increases safe area and lowers out-put capacitance but increases r_p and saturation resistance, however, it is difficult to grow specific high resistivity material to a high yield which may again increase costs; 3 - The base width may be increased to expand safe area but gain, frequency cutoff, and r_p are reduced; 4 - Several constructional features have been reported as increasing safe area without compromise - making uniformly parallel junctions with germanium alloy types, avoiding pipes in the regrowth area of germanium indium alloy types, avoiding narrow base-emitter spacing and extremely sharp corners with the interlocking teeth of a silicon annular inter-digitated device. (This may increase the amount of silicon required to insure sufficient emitter periphery for good high current gain)

Let us examine these conditions with some actual examples. The graph of Figure 5-8 shows the 40 volt dc safe power limit (based on secondary breakdown failure) for three different junction areas of germanium alloy transistors. All transistors were made with 8-12 ohm cm material, had dot emitters, base width of 1 to 3 mils all used indium as the collector alloy, and were at least 75 volts, BVCEs. As is expected, the collector area definitely affects the safe area.

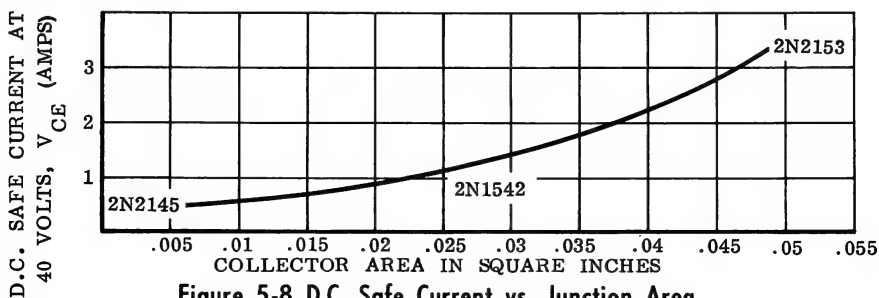


Figure 5-8 D.C. Safe Current vs. Junction Area

Increased resistivity affects safe area mainly due to increased avalanche voltage. For example, in Figure 5-9 the safe area voltage for a 500 μ sec pulse is shown at 500 mA and 5 amperes plotted against resistivity. The graph indicates that resistivity effects the safe area at low currents more than it does at high currents. These curves are not to be construed as defining breakdowns vs. resistivity, but are approximate only.

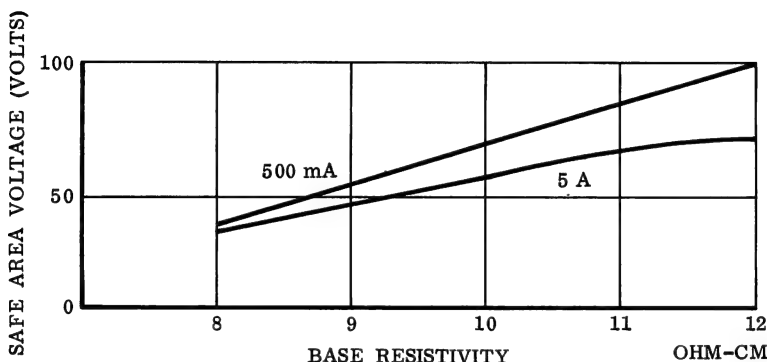


Figure 5-9 500 μ sec Safe Voltage at 0.5 A & 5A vs. Base Resistivity

The exact effects of base-width are more difficult to plot since no devices are available with a known base width vs. safe areas with all other factors equal. However, we can plot two devices both with similar resistivity and area but with radically different base widths. Unfortunately there is radically different material used and methods employed in their manufacture. The 2-3 mil base-width device is an aluminum-doped indium germanium alloy device with a step junction impurity profile whereas the device with 0.2 mil base-width uses lead-antimony employing a push-out base alloy-diffused construction method which gives a graded impurity profile. The dc safe areas of these devices are compared in Figure 5 - 10.

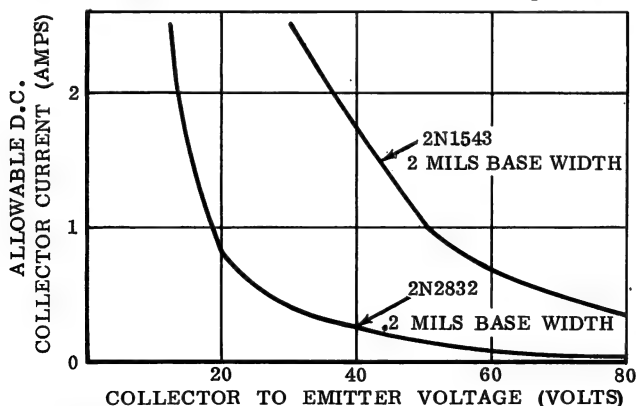


Figure 5-10 Effect of Base Width on D.C. Safe Area

As a blow to our cherished hopes of a universal transistor, there apparently is some law which demands that dc safe area multiplied by the frequency cutoff equals some constant. A high frequency device which is optimized for 50 Mc power gain will have poor dc safe area, and a low frequency device optimized to have high dc safe area will generally have low frequency cutoff. Obviously, the end use dictates which transistor to use. Thus, power supply designers need a transistor with a large dc safe area for the series pass regulator and must sacrifice frequency response with present day devices. However, a designer of RF amplifiers or of switching inverters wants a very high frequency device with a safe area (for a few microseconds) as large as possible, and could care less what dc safe area is. Fortunately, there exists both germanium and silicon devices which fulfill both specifications.

Thus, this has led Motorola and other manufacturers to bring out special lines of devices. For example, the Motorola RF amplifier line of annular silicon devices (2N2947-2952) and the much older 2N1529-1560 series of low frequency germanium devices intended for power supply and low frequency inverter use.

Unfortunately, this leaves a hole in the use pattern for a wide band device which could be used in hi-fi requirements and other uses. To fill this socket a somewhat modest frequency device is generally used (about a 1 Mc) with reasonably good dc safe area and feedback to enhance the upper frequency performance.

It should be obvious by now, that, data sheets must include safe area information (from dc to fast pulses) as well as power dissipation capabilities so as not to mislead design engineers. There is presently a trend to do this and is being pushed by the JEDEC committees on power transistors as well as EIA. Many new devices will have safe area curves or at least some information concerning secondary breakdown registered in the 2N number format. Older devices will have safe area design curves added to the data sheets although the information will not be registered and the option of doing this or not will be up to the transistor manufacturer. Motorola has taken the position to release as much of this information as possible via Application Note # 137 and adding curves to data sheets as they are revised.

Meanwhile back in the Lab - the search for the "Universal Transistor" continues and who knows, perhaps one day it will be found.

SECTION 6

Motor Control

GENERAL

Although motor control is commonly thought of as speed control, many other useful motor control functions such as switching, direction control or programming may be performed by semiconductor devices.

Speed control may be most easily accomplished on motors whose speed is a function of power input. This is generally true of dc or universal motors. In most types of ac motors, however, speed control must be accomplished by varying the frequency. A variable frequency drive may be made with an inverter such as the one described in Section 7-4, in conjunction with a variable frequency trigger. In some cases, the control range is limited because of rapid fall off in torque as frequency is reduced.

Speed control by variation of average power input to a dc motor is accomplished by pulse width modulation; that is, by switching between the "on" state and the "off" state and controlling the percentage of "on" time. In ac circuits, this is rather simply accomplished by phase control; that is, by controlling the phase relationship between the gate trigger signal and the supply voltage, the percentage of "on" time in each cycle can be controlled. In a dc circuit, such as a parallel inverter, pulse width modulation may be accomplished by use of a constant repetition rate and a variable time delay between the turn-on pulse and the turn-off pulse, thereby, permitting adjustment of the on time in each cycle. Alternatively, similar results may be obtained by using a fixed time delay between the on pulse and the off pulse, and varying the repetition rate. This will produce an output with a fixed on time and a variable off time, permitting adjustment of the percentage of on time.

SECTION 6-1

Phase Control with Feedback

One of the drawbacks of motor speed control by variation of the average power input is that as the power input is reduced to reduce the speed, the available torque is also reduced. This disadvantage may be overcome by using a feedback signal to advance the firing angle in proportion to the load on the motor; thereby, supplying it with a greater amount of power as more torque is required.

The circuit in Figure 6-1-1 derives its feedback from the load current and does not require separate connections to the motor field and armature windings. Therefore, this circuit can be conveniently built into an appliance or used in a control box separated from the appliance.

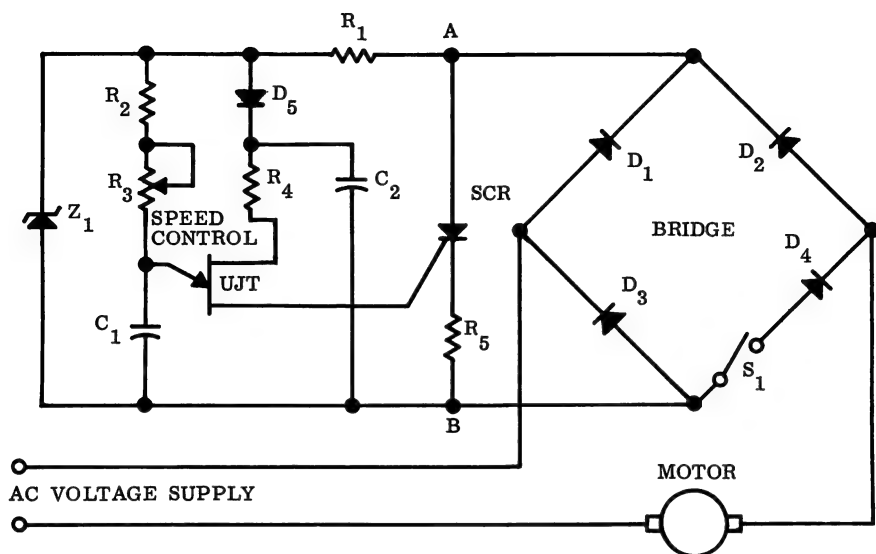


Figure 6-1-1 SCR Feedback Motor Speed Control

TABLE 6-1-1

PARTS LIST FOR FIGURE 6-1-1

SCR Motorola MCR 808-4	R ₁	18K, 2W
UJT 2N3480	R ₂	3.9K, 1/2W
D ₁ , D ₂ 1N3493 -- 200V, 18A Rectifier	R ₃	50K, Potentiometer
D ₃ , D ₄ 1N3493R -- 200V, 18A Reverse Polarity Rectifier	R ₄	330 Ω , 1/2W
	R ₅	See table
D ₅ 1N4001 -- 50V, 1A Rectifier	C ₁	.1 μ fd
Z ₁ 1N751A -- 5.1V, 400 mW Zener Diode	C ₂	10 μ fd, 10 V

NOMINAL R ₅ VALUES			
MOTOR RATING (AMPS)	R ₅		$R_5 = \frac{2}{I_M}$ $I_M = \text{Max Rated Motor Current (RMS)}$
	(OHMS)	(WATTS)	
2	1	5	
3	0.67	10	
6.5	0.32	15	

The circuit operates as follows: Diodes D₁, D₂, D₃, and D₄ form a bridge which applies a full-wave rectified voltage to the SCR. When the SCR conducts, the bridge appears as a very low resistance in series with the motor. The normal line voltage, less the drop across two of the rectifiers, the SCR and R₅, is applied to the motor. By delaying the firing of the SCR until a later portion of the cycle, the voltage applied to the motor is reduced and its speed is reduced proportionally.

Phase delay of the SCR firing is obtained by the charging of C₁ through resistors R₂, and R₃ from the voltage level established by the zener diode, Z₁. When C₁ charges to the firing voltage of the unijunction transistor, it conducts triggering the SCR. As C₁ discharges, the emitter current of the unijunction transistor drops below the holding current level and the unijunction ceases to conduct. While the SCR is conducting, the voltage drop between points A and B falls below the breakdown voltage of Z₁. Therefore, during the conduction period, the charging of C₁ is controlled by the voltage drop from A to B and resistors R₁, R₂, and R₃. Since the voltage between A and B is a function of motor current, C₁ is charged during the conduction period to a value which is proportional to motor current. The value of R₅ is chosen such that C₁ cannot charge to a high enough voltage to fire the unijunction during the conduction period. However, the amount of charging required to fire the unijunction transistor has been decreased by an amount proportional to the motor current. Therefore, the firing angle at which the unijunction transistor will fire has been advanced in proportion to the motor current. As the motor is loaded and draws more current, the firing angle of the unijunction transistor is advanced even more, causing a proportionate increase in the voltage applied to the motor, and a consequent increase in its available torque.

Since the firing voltage of the unijunction transistor depends upon the Base 2 to Base 1 voltage, it is necessary to support the Base 2 voltage during the conduction portion of the cycle to prevent the feedback voltage from firing the unijunction. D₅ and C₂ perform this function.

Nominal values for R₅ can be obtained from the table or they can be calculated from the equation given. Exact values for R₅ depend somewhat on the motor characteristics. Therefore, it is suggested that R₅ be an adjustable wire-wound rheostat which can be calibrated in terms of motor current, and the speed control can be adapted to many different motors. If the value

of R_5 is too high, feedback will be excessive and surging or loss of control will result. If the value is too low, a loss of torque will result. Since R_5 must carry the maximum motor current, its wattage is determined accordingly.

This circuit has been operated successfully with 2 and 3 ampere 1/4 inch drills and with a 6.5 ampere 1/2 inch drill and has satisfactorily controlled motor speeds down to 1/3 or less of maximum speed with excellent torque characteristics.

To demonstrate the improvement of performance which is possible with feedback, speed-torque curves were run on the 6.5 ampere 1/2 inch drill. The curves are shown in Figure 6-1-2. The marked improvement in torque when feedback is employed is readily seen. If sufficient loading capability were available, the curves could have been made to intersect the full power speed-torque characteristic. Note that with a .35 ohm feedback resistor (R_5) at 60° conduction angle, there is a slightly rising speed-torque characteristic. This is an undesirable condition since it causes oscillation. By switching to half-wave operation, the speed at which this rising characteristic occurs can be approximately halved.

If lower cost is desired, a circuit such as that shown in Figure 6-1-3 may be used. This type of circuit derives a feedback signal from the armature counter EMF which is produced by the residual field.

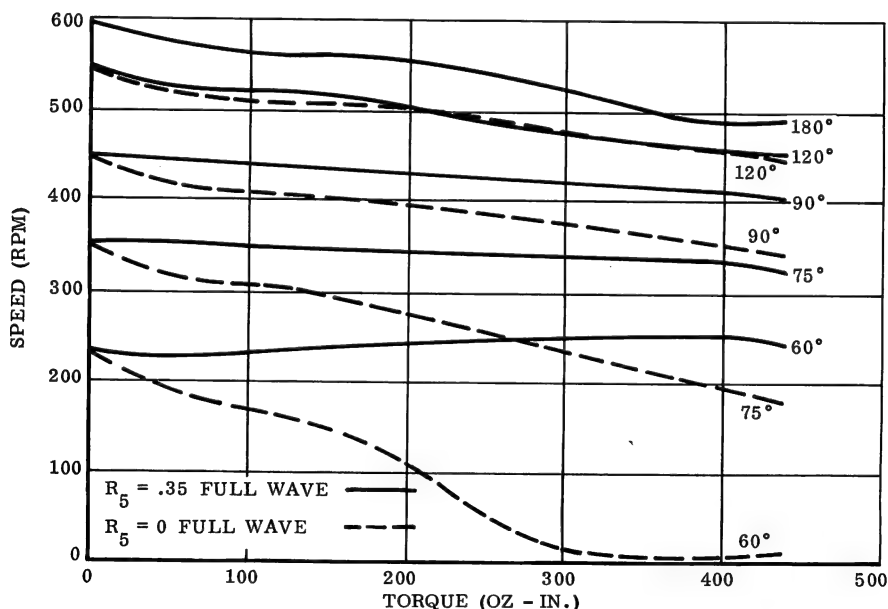


Figure 6-1-2 Speed vs. Torque for Circuit of Figure 6-1-1

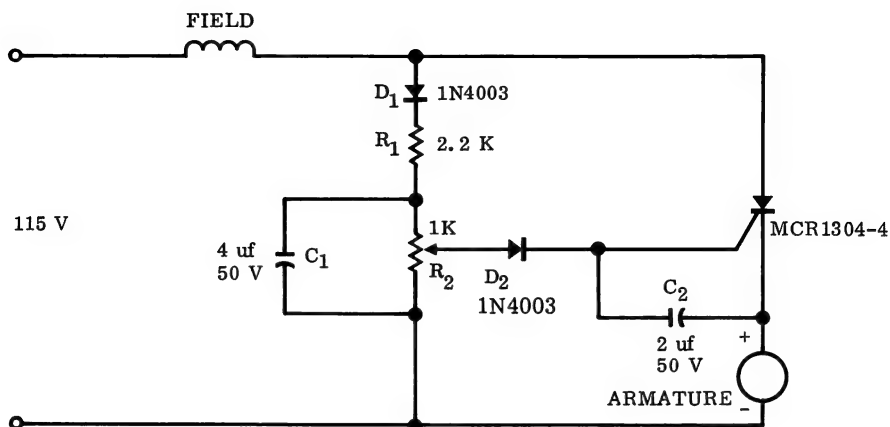


Figure 6-1-3 Speed Control with Counter EMF Feedback

In the circuit shown, firing is accomplished when the voltage on the wiper arm of potentiometer R_2 rises to a high enough value to forward bias diode D_2 , allowing gate current to flow. Since the counter EMF tends to reverse bias D_2 , the firing point depends upon the counter EMF. The counter EMF in turn, is a function of speed. As the motor is loaded reducing the speed and thereby, the counter EMF, diode D_2 becomes forward biased earlier in the cycle; thereby triggering on the SCR earlier in the cycle and supplying the motor with greater power to offset the effect of loading.

In general, for universal motors, the feedback voltage will be low, because the residual field is low. Usually it is found to be in the 1-5 volt range. For maximum feedback effectiveness, the change in counter EMF between zero and full load on the motor should approach the peak to peak voltage of the ramp function generated by the trigger circuit. Because of the voltages involved, low cost breakover type devices are ruled out and the trigger circuit is pretty much limited to the type of circuit shown in Figure 6-1-3.

If the field can be kept at a high value during the off time of the SCR, such as would be the case with a permanent magnet motor, the same technique could be applied to a breakover device type of trigger circuit.

The circuit values shown in Figure 6-1-3 are representative. For optimum performance they should be adjusted for the motor with which they will be used.

SECTION 6-2

Chopper Type Speed Control

Figure 6-2-1 shows a chopper type dc motor drive. SCR₁ is the load SCR and must be capable of carrying the maximum desired load current. The function of SCR₂ is to commutate SCR₁. Assuming that we wish to design a speed control circuit to perform a function similar to that described in Section 6-4, SCR₁ would have to be capable of carrying 200 amperes. This might be accomplished economically by parallel operation of a low cost SCR such as the MCR808 or the MCR1304. Of course, the necessary precautions for current sharing and triggering would have to be taken. The value of the commutating capacitor C_C may be calculated approximately from the following equation:

$$C_C \approx 1.4 \frac{t_o I_A}{V_C},$$

where t_o is the turn-off time of the device under the conditions of operation, I_A is the anode current just before commutation, and V_C is the voltage across C_C just before the start of commutation.

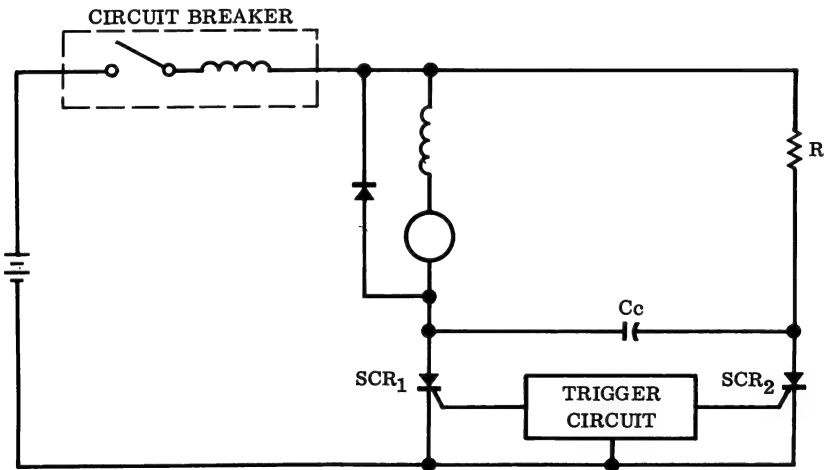


Figure 6-2-1 Speed Control with Resistive Charging

If SCR₁ is a parallel combination of small devices, rather than a single large device, there will be a decrease in the required value of CC. The required value might be decreased by 50 per cent, for example, over a single high current SCR based on a turn-off time for the MCR1304, of 16 microseconds. For a current of 200 amperes, and a supply voltage of 36 volts, the commutating capacitor required is approximately 90 microfarads. The minimum value of R depends upon the minimum time required to charge CC. Since a repetition rate of 30 cycles per second was chosen, along with a minimum on duty cycle of 10 per cent, the minimum charge time would be 3.33 milliseconds. Setting this time equal to three time constants, we obtain

$$3 R(90 \times 10^{-6}) = 3.33 \times 10^{-3}$$

$$R \approx 12 \text{ ohms}$$

The dissipation in R based on a worst-case 90 per cent duty cycle would be approximately 108 watts. Another approach that will permit the virtual elimination of that dissipation as well as a further reduction of CC is illustrated in Figure 6-2-2 below.

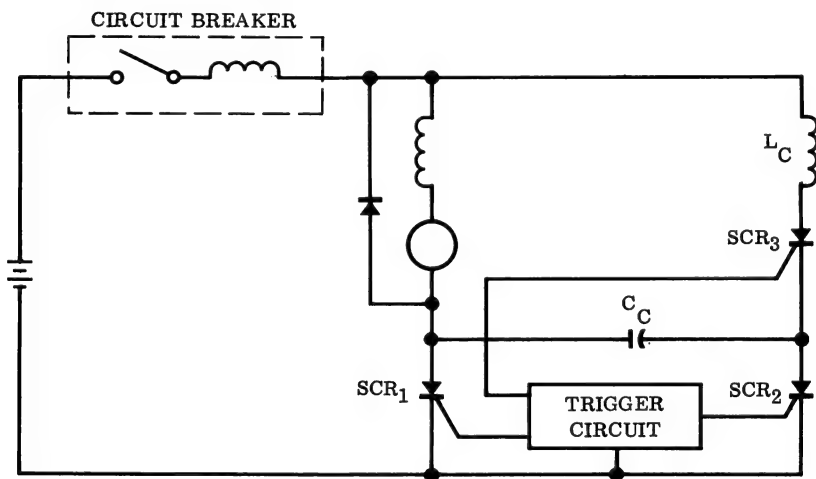


Figure 6-2-2 Speed Control with Inductive Charging

In this circuit, SCR₃ which is gated on at the same time as SCR₁ allows the resonant charging of CC by LC to twice the supply voltage. This permits the reduction of CC by a factor of 2. In high current applications, the saving in the cost of the commutation capacitor can more than offset the cost of the additional SCR. The value of L_C determines the charging rate and charging current for C_C. The maximum value of L would be based on a resonant frequency whose half cycle time is equal to 3.33 milliseconds, or 150 cycles. With a commutating capacitor of 45 microfarads, the inductance would be approximately 25 millihenries. It should be remembered,

of course, that with the commutating capacitor charged to twice the supply voltage, the SCR must have a reverse voltage capability of that same amount. It would be desirable to use a smaller value of inductance and decrease the charging time. It should be kept in mind, however, that as the charging time decreases, the peak charging current increases. Also, in order to charge to twice the supply voltage, the Q must be kept high.

Comparing the SCR motor control with the transistor version, it is found that the transistor version has a lower voltage drop. At 200 amperes, the drop across the transistor control would be about 1.3 volts; whereas that across the SCR version would be about 1.5 volts. The losses in the SCR control are about 33 per cent resistive whereas those in the transistor control are approximately 75 per cent resistive. Therefore, the loss would drop off more rapidly as the current decreases, in the transistor control than they would in the SCR control. On the other hand, the SCR control could be made lower in cost. Moreover, it is more resistive to damage by fault conditions than is the transistor control. A short across the motor would cause the current to rise too rapidly for the current limiting feature to save the transistors. However, that sort of overload could be withstood by the SCR control, until the circuit breaker has tripped.

SECTION 6- 3

SCR Windshield Wiper Control

At times it is necessary to control the direction and/or time of operation of a motor as well as the speed. An example of such an application is the automatic washing machine. An SCR control could periodically reverse a motor which directly drives the agitator to give the required agitation cycle without the necessity of a mechanical transmission. It could then be easily switched to high-speed single-direction operation for the spin-dry cycle. Another example of programmed operation would be an automatic tapping machine. Best results are obtained when the tap is advanced several turns and then partially withdrawn so that it can clear. Speed adjustment is required for different taps and material. All this could be easily combined into a single SCR control.

Perhaps a simpler example of programmed operation can be found in the automotive field. It is often desired to operate automobile windshield wipers at reduced speed when precipitation is light. However, it is annoying to the driver to have the wiper blade move slowly across the windshield and perhaps stall part way across. He would prefer to have the blade move rapidly across the windshield at full wiping speed, but less frequently. The simple circuit of Figure 6-3-1 accomplishes this purpose. The unijunction trigger circuit is designed to give a broad range of repetition rates. The fastest represents continuous wiping action and the slowest represents the lowest desired wiping frequency. When the SCR is triggered on the motor begins to operate wipers and also closes S_2 . S_2 short circuits the SCR, turning it off, and carries the motor current itself. S_2 remains closed for one complete wiping cycle, at which time it opens. Since the SCR has previously been turned off, the wiper will remain in the rest position until another trigger pulse from the unijunction initiates another wiping cycle. Capacitor C_2 protects the SCR and the contacts of S_2 from the voltage transient due to the opening of the motor circuit. The circuit was designed as shown in Figure 6-3-1 with the SCR connected in the positive leg of the wiper motor, so that it might accommodate the most common situation encountered today. That is a negative ground electrical system, with the shell of the motor connected to the car frame and serving as the negative terminal. For new designs, however, it would be more desirable to bring out a separate negative terminal so that the SCR could be connected in the negative leg of the motor permitting the elimination of transformer T_1 by directly connecting base 1 of the UJT to the gate of the SCR. Of course, S_2 would also have to be in the negative leg of the motor.

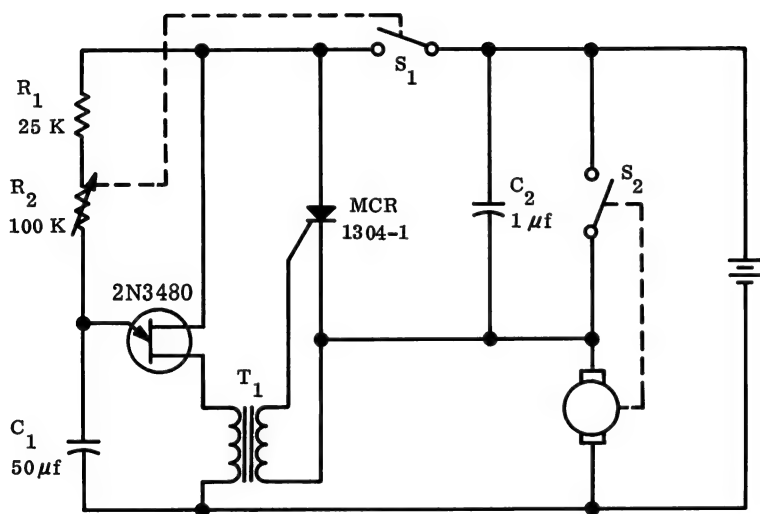


Figure 6-3-1 SCR Windshield Wiper Control

SECTION 6-4

Pulse Width Modulation DC Motor Control

In order to control the speed of a dc series field motor at different required torque levels, it is necessary to adjust the voltage applied to the motor. For any particular constant voltage the motor speed is determined solely by the torque requirements and top speed is reached under minimum torque conditions. When a series motor is used as a traction drive for vehicles, it is desirable to control the voltage to the motor to fit the various torque requirements of grades, speed and load. The common method of varying the speed of the motor is by inserting resistance in series with the motor to reduce the power supplied. This type of motor speed control is very inefficient and wasteful of battery power, especially under high-current, high-torque conditions due to the I^2R loss.

A more efficient method of controlling the voltage applied to the motor is the pulse-width modulation method shown in Figure 6-4-1. In this method, a variable width pulse of voltage is applied to the motor to proportionally vary the average voltage applied to the motor. A diode is placed in parallel with the inductive motor path to provide a circuit for the inductive motor current when the switch is opened. This circulating current path prevents abrupt current changes and resultant high voltage across the switching device. The current decreases only in response to motor and diode loss.

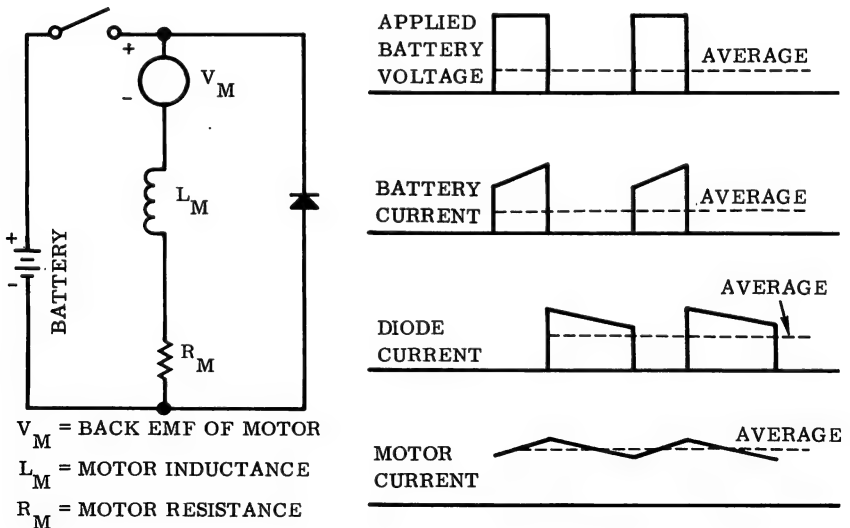


Figure 6-4-1 Basic Pulse Width Modulated Motor Speed Control

With reference to Figure 6-4-1, it can be seen that the shunt diode will allow more average current to flow through the motor than is taken from the battery. However, the power taken from the battery is approximately equal to the power delivered to the motor indicating that the energy stored in the motor inductance at battery voltage level is delivered to the motor at the same approximate current level but at a much lower voltage level when the battery is disconnected.

To provide smooth and quiet motor operation, the current variations through the motor should be kept to a minimum during the switching cycle. There are limitations on the amount of energy that can be stored in the motor inductance, which, in turn, limits the power delivered to the motor during the off time; thus, the off time must be short. For low speeds the on time must be short and a rapid switching rate is required which rules out mechanical means. Practical solutions can be found in the use of solid state devices for fast, reliable and efficient switching operations.

POWER TRANSISTOR MOTOR SWITCH

Present day series motors operating as tractive motors require high current levels depending upon application and condition of operation. For fork lifts, golf carts and other small electric vehicle applications, a pulse-width modulated motor control may be required to handle current levels in the vicinity of 400-500 amperes unless current limiting circuitry is used. Even with current limiting circuitry, torque demands require switching currents as high as 250-300 amperes. This high-current demand is not met by any single power transistor at the present time and units must be paralleled to meet this current level.

The maximum current level that any transistor can handle in any particular circuit is dependent upon the switching load line and the safe area of operation. (See Section 5.) For the pulse-width modulator there is a possibility of two different load lines depending upon whether a capacitor is used to tailor the load line as shown in Figure 6-4-2. The decision to be made is to either parallel the motor with the high capacitance necessary to tailor the load line or add additional transistors in parallel to meet the rectangular load line safe area requirements. One additional factor is that with each transistor operating at a lower current level and hence, higher gain, the total drive current requirements for the transistor switch is generally reduced.

Because of the variation in $V_{BE(sat)}$ for each transistor, collector current levels of each transistor in the parallel combination can be quite different unless some precautions are taken to equalize the drive requirement. The recommended method is to insert a resistance in series with each transistor emitter. The effect of this emitter resistor is shown graphically in Figure 6-4-3. The variation of collector current that is acceptable determines the value of emitter resistance required. This value of emitter resistance can be reduced by pre-selecting $V_{BE(sat)}$ to a closer tolerance.

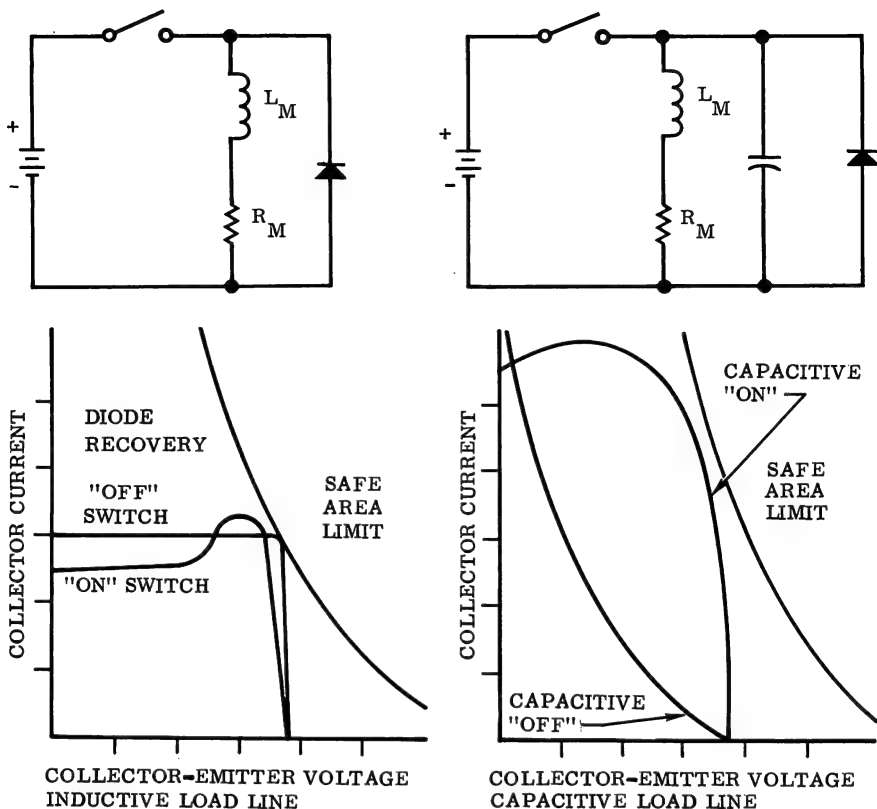


Figure 6-4-2 Safe Area and Load Line Considerations

DRIVE CIRCUITS

Power loss in the drive circuit can be quite high as illustrated in the drive circuits of Figure 6-4-4. The 36 volt, 200 ampere motor requirement determines a drive current of at least 10 amperes, assuming force gain of 20, which would dissipate 350 watts in the collector resistor of the emitter follower driver. If the saturated Darlington circuit is used, the drive current contributes to the load current. However, the .5 volt drop across the saturation resistor in series with the collector can dissipate 95 watts at the maximum current level of the motor. Of course, the power loss in this resistance will be in direct proportion to the peak motor current demand while the peak power dissipation in the emitter follower circuit will be a constant and independent of motor demand.

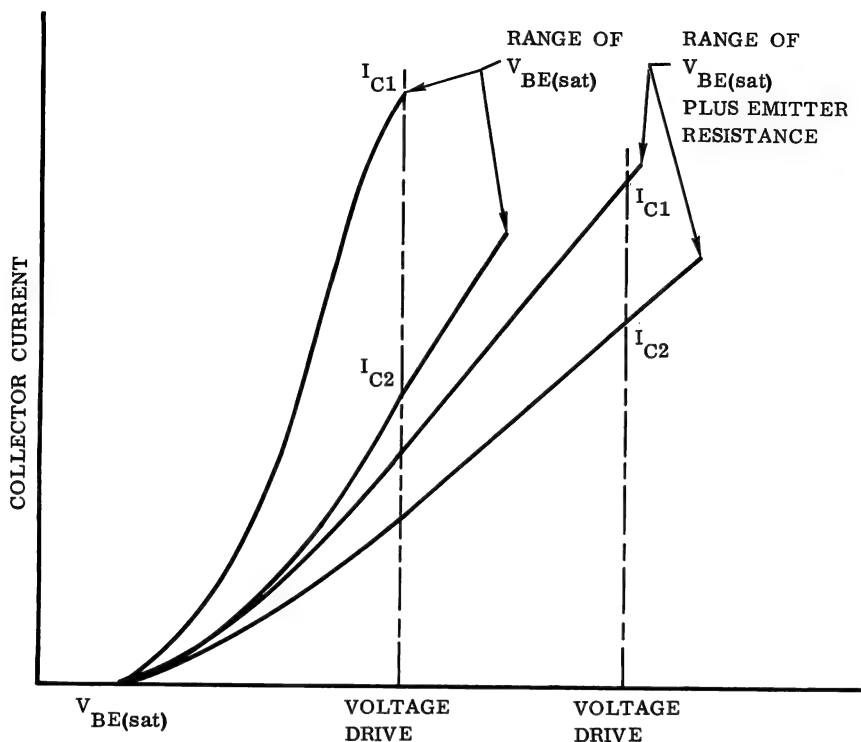


Figure 6-4-3 Effects of Emitter Resistance on Parallel Drive

The value of the saturation resistor is determined from $V_{BE(sat)}$ and $V_{CE(sat)}$ information obtained from the device characteristic curves with the following relationship:

$$R_{(sat)} = \frac{V_{BE2(sat)} + V_{CE1(sat)} - V_{CE2(sat)}}{I_L}$$

With reference to Figure 6-4-4:

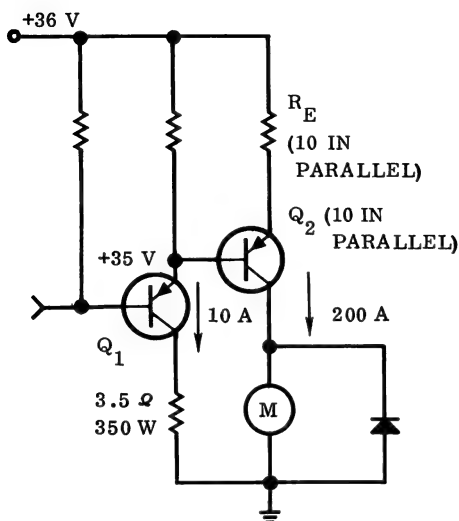
$R_{(sat)}$ = saturation resistor in series with collector of Q_2

$V_{BE2(sat)}$ = saturated emitter-base voltage of Q_2

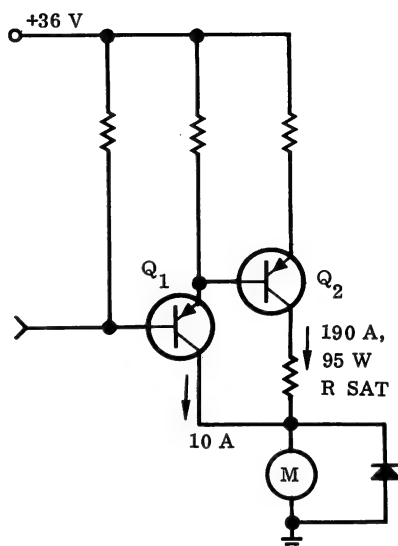
$V_{CE2(sat)}$ = saturated collector-emitter voltage of Q_2

$V_{CE1(sat)}$ = saturated collector-emitter voltage of Q_1

I_L = load current



EMITTER FOLLOWER DRIVE



SATURATED DARLINGTON DRIVE

Figure 6-4-4 Drive Circuits

It can also be shown that $V_{R(sat)} \approx V_{BE1(sat)}$ where $V_{R(sat)}$ = voltage drop across saturation resistor of Q_2 .

Because the drive transistor in the Darlington circuit will begin to turn on before the final transistor, it is apparent that unless the input to the driver is restricted, this transistor may switch a load line that is beyond its safe area. A driver transistor should be selected with a safe area that extends to a collector current and battery voltage level, that will not be reached by available input to the driver. Input characteristics that taper and do not apply full input to the driver until the finals have turned on, will also keep the driver within its safe area.

Because most high current transistors have slow switching characteristics, they tend to match up in load line excursions much closer than the faster switching transistors. However, slow switching transistor power dissipation is greater and if switching rate is rapid, reverse bias may be necessary to reduce switching power dissipation by decreasing "switching off" time. This reverse bias can be produced by a separate transistor converter circuit or by charging a capacitor during the on drive and discharging it in the reverse direction to drive "off" the output transistors. A check should be made to determine what mismatch can be expected in respective storage and switching times and the resultant load line differences.

PULSE-WIDTH OSCILLATOR

One of the most useful circuits for producing a variable pulse-width at a constant switching rate is shown in basic form in Figure 6-4-5. The

unijunction oscillator (Q_1) is charged through the potentiometer (R_{11}) and that side of the flip-flop which is in the off condition (Q_2 or Q_3). When capacitor (C_1) is charged to a voltage level where the emitter voltage reaches the peak point voltage, the emitter to B_1 junction of the unijunction will turn on and supply a trigger to the flip-flop. The "off side" of the flip-flop will turn "on" and the "on side" will in turn be turned "off." The capacitor will now charge from the opposite side of the potentiometer. This charging of the unijunction capacitor from alternate sides of the potentiometer can give two different time constants which, in turn, determines the "on" and "off" time for whichever side of the flip-flop is determined to be the drive for the next stage. The potentiometer will give a wide range of pulse-width control at a fairly constant switching rate over a wide temperature range.

Silicon diodes, D_3 and D_4 are necessary to disconnect the charging potentiometer from the voltage divider current that could flow through the transistor that is conducting. Silicon diodes D_2 and D_5 are necessary to eliminate the effect of transistor bias current levels on the charging current through potentiometer R_{11} . These last two diodes are sometimes not needed but are recommended for stable operation.

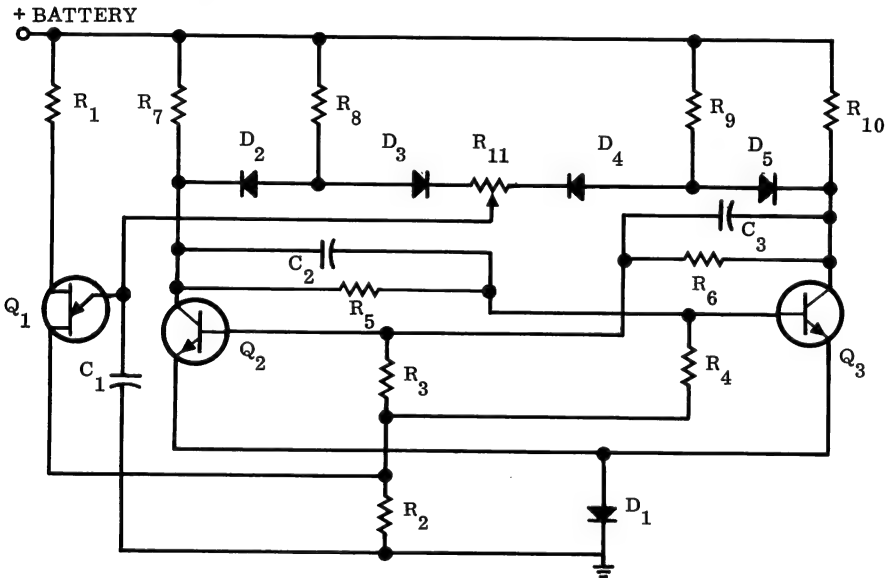


Figure 6-4-5 Basic Variable Pulse-Width Oscillator

CURRENT LIMITING

Because of the high motor current levels that are encountered when operating under stall or starting conditions, a current limiting circuit becomes desirable. The motor current should be limited to a value that will only meet maximum torque requirements to reduce high motor temperatures and total number of parallel power transistors. This current limiter

would sense the total current through the transistors and when a preset current level is reached, the drive circuitry is turned off. The current can be sensed by the drop across a resistor common to all transistors or sense the voltage drop across an individual transistor emitter resistor in a well matched parallel combination. The limiting should be as near instantaneous as possible and be repeatable at the preset limit under all conditions of operation.

The current limiting circuit of Figure 6-4-6 is excellent for this application and is discussed in detail in Section 8-2. The current level at which the transistor will switch on is determined by the calibration to the basic characteristics of the tunnel diode as shown in Figure 6-4-6.

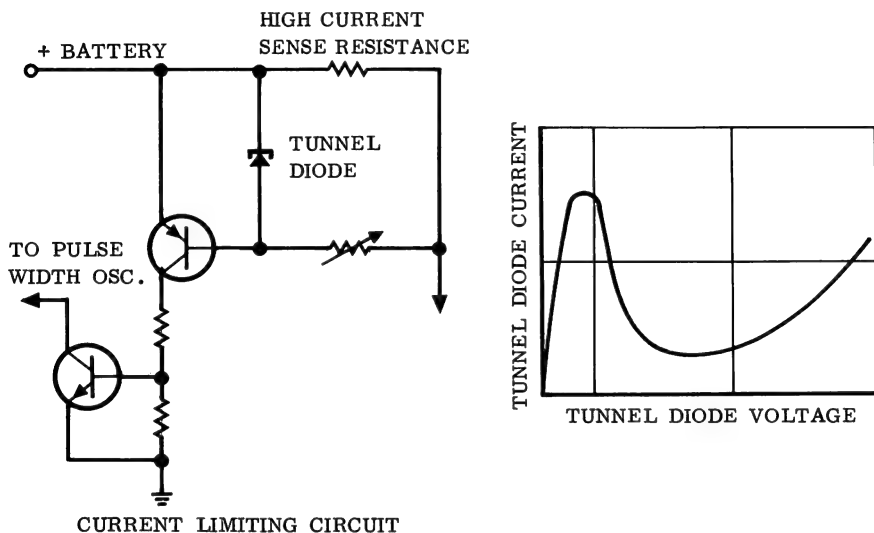


Figure 6-4-6 Basic Current Limiting Circuit and Tunnel Diode Characteristics

COMPLETE MOTOR CONTROL CIRCUIT

A complete circuit for a simple pulse-width modulated series motor control is shown in Figure 6-4-7. This circuit was built to operate the motor in a Cushman golf cart which operates from 36 volts. The current level through the motor approaches 300 amperes under stall or starting conditions. It was found that current limiting had to be no less than 200 amperes in order to have torque enough to climb steep inclines. Therefore, a minimum of (10) MP505 or 2N1165 transistors in parallel are required if capacitor tailoring of the load line is not used.

If current limiting action is sustained for any length of time, the power transistor can attain a high temperature unless the 300 cps limiting frequency is reduced or reverse bias is used to decrease switching power. The 300 cps limiting frequency can be reduced by increasing the minimum time of

recharging the unijunction capacitor or employing a monostable multivibrator in the sense circuit to control the off time of the pulse-width modulator circuitry when limiting.

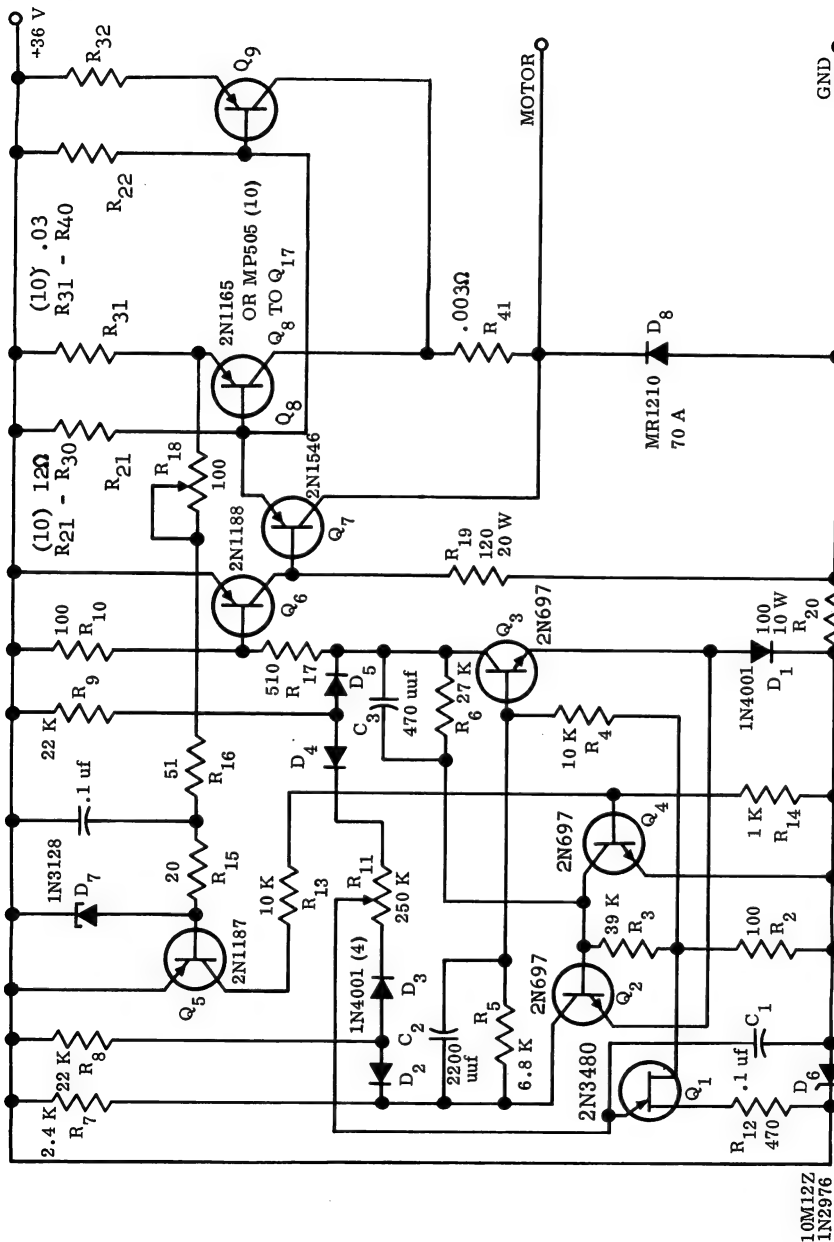


Figure 6-4-7 Complete Pulse-Width Modulated Motor Control (36V - 200 A)

SECTION 6-5

Commutatorless DC Motor

In many portable equipment applications, hazardous locations are encountered where the arcing of dc motor brushes cannot be tolerated. The elimination of contacts and brushes can also increase the reliable operating life of the motor. It is apparent that if a transistorized dc to ac inverter were available, a brushless ac induction motor could be used for these applications. However, it is also apparent that a separate transformer is not necessary because feedback windings can be added to the regular induction motor windings to operate a push-pull transistor inverter circuit. The transistor operated commutatorless motor circuit shown in Figure 6-5-1, illustrates this basic circuit configuration.

One important factor that must be considered in the design of this type of inverter circuit is the highly inductive load line which will be switched by the transistor due to the high leakage inductance in the motor. This inductive load line will generally require collector-emitter zener protection to protect against high voltage stresses.

The frequency at which the circuit oscillates will determine the speed of the motor. If the motor iron used has a poor saturating BH curve, adjusting the base drive resistance can have some control of frequency while still operating in the saturated iron mode for the push-pull transistor inverter. If the based drive is decreased to a point where the iron does not tend to saturate to determine oscillator frequency, the frequency will change greatly with motor load requirements.

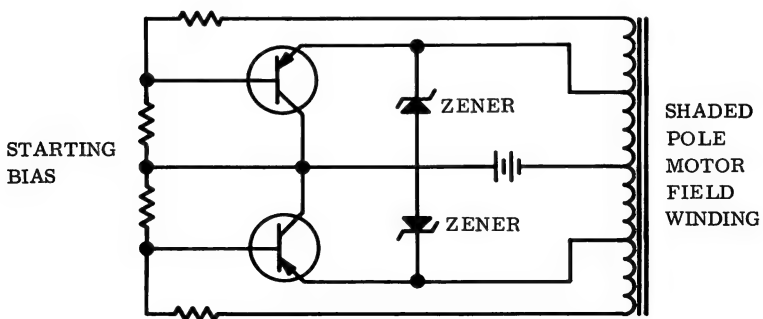


Figure 6-5-1 Basic Commutatorless Motor



SECTION 7

Inverters

SECTION 7-1

Low Voltage Inverters and Converters

There are applications where low dc voltage sources such as solar, thermoelectric, fuel cells, and many other low voltage sources must be converted to a higher voltage. For these applications, inverters and converters can be utilized to change the voltage levels.

To obtain good efficiency when operating inverters or converters from low voltage sources, the transistors used must have low saturation voltages. Several transistors designed for low voltage inverter service have been introduced in the past year.

A LOW-VOLTAGE, HIGH CURRENT INVERTER

This section discusses an inverter circuit that changes a 2 volt dc source to a higher voltage square-wave. The extremely low saturation voltage of the 2N2728 power transistor enables switching currents up to 50 amperes efficiently. The saturation voltage of the 2N2728 transistor is a maximum of 0.1 volt at 50 amperes of collector current when driven with a forced gain of 10.

CIRCUIT DESCRIPTION

The circuit, shown in Figure 7-1-1, is a two transformer dc to ac inverter which can easily be changed to a dc to dc converter by rectifying and filtering the output. The two transformer inverter circuit was used instead of a conventional one transformer inverter to minimize core losses in the output transformer where high collector currents are switched. In a single transformer inverter, high peak currents are encountered during the time the core saturates.

The driver transformer utilizes 6 mil Hi Nickel Steel laminations for the core. The core consists of 1/4 inch wide laminations stacked one inch high. The frequency of oscillation is determined by the following equation:

$$f = \frac{V_{FB} \times 10^8}{4N_4AB_S}$$

A = the cross sectional area of the core in square centimeters.

B = the saturated flux density in lines per square centimeters.

N_4 = the number of primary turns.

V_{FB} = the peak feedback voltage across the primary in volts.

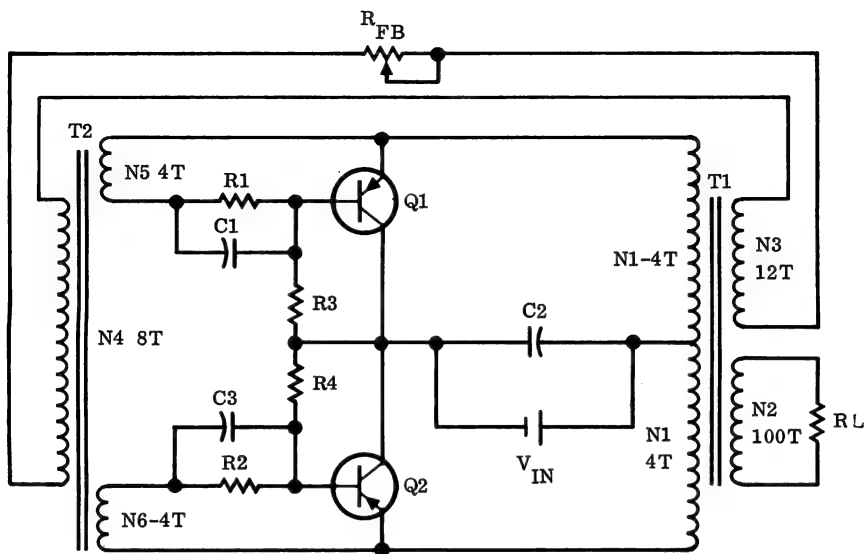


Figure 7-1-1 Low Voltage High Current Inverter

TABLE 7-1-1

Parts List for Figure 7-1-1

R_1	0.75Ω , 5 W
R_2	0.75Ω , 5 W
R_3	7.5Ω , 5 W
R_4	7.5Ω , 5 W
R_{FB}	1Ω , 5 W
C_1	$20\mu f$, 6 V
C_2	$10,000\mu f$, 6 V
C_3	$20\mu f$, 6 V
T_1	Phoenix Transformer PX 2127
T_2	Phoenix Transformer PX 2126
Q_1, Q_2	2N2728
V_{in}	2 V, 50 A

The frequency is set by adjusting the feedback voltage. For this application, the core was set to oscillate at 1 kc. A schematic representation of the driver transformer is shown in Figure 7-1-2.

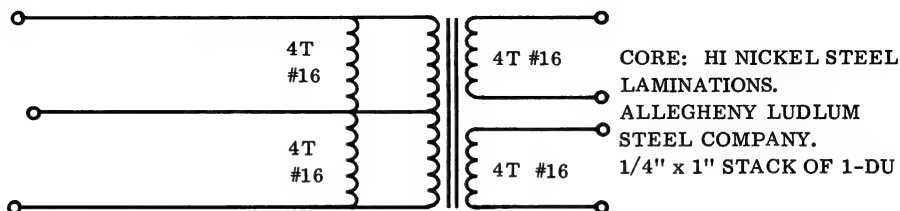


Figure 7-1-2 Driver Transformer PX-2126

The output transformer was wound on two 12 mil Silectron "C" cores (Arnold AA-86). This transformer was designed to oscillate at 850 cycles per second to insure non-saturated operation at 1 kc. To keep the dc resistance of the primary windings at a minimum, parallel primary windings were used.

The circuit operates in the common emitter configuration to take advantage of the high current gain of the 2N2728 transistors. The secondary of the driver transformer was split to allow driving each transistor separately. By driving each transistor separately, the collectors could be tied together by mounting to a common heat sink, without insulating the case of the transistor from the heat sink; thereby minimizing the case to heat sink thermal resistance. Both transistors were mounted on a 3" x 1" x 1/8" copper plate and the negative terminal of the battery was bolted to this copper plate, to minimize wiring resistance.

The resistors R_1 through R_4 are starting resistors. These resistors function as a voltage divider and provide a starting bias to the base.

Capacitors C_1 and C_3 were used to enhance switching speed.

CIRCUIT PERFORMANCE

The circuit was set to oscillate at approximately one kilocycle per second by adjusting the feedback voltage by means of a resistor in the feedback winding. The value of this resistor was approximately 1/2 ohm.

The circuit operates at maximum efficiency of greater than 70 per cent at 60 watts output, as shown in Figure 7-1-4. As the circuit load is increased the output voltage decreases because the transistors do not saturate.

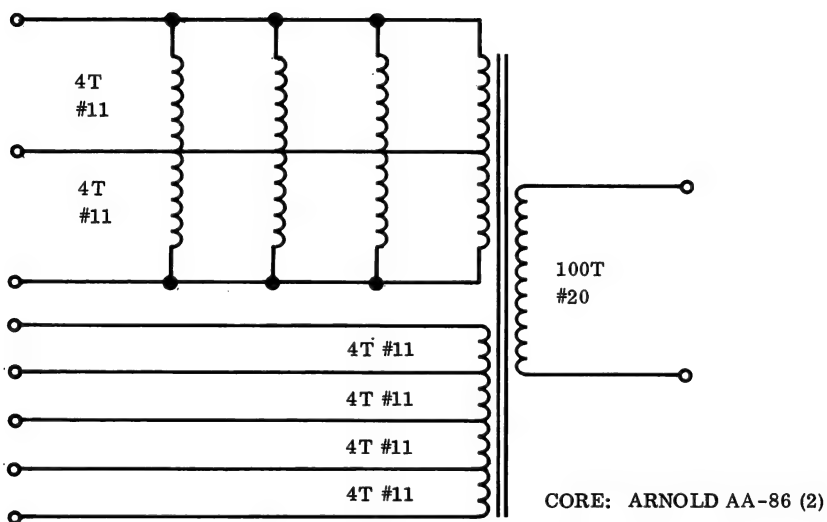


Figure 7-1-3 Output Transformer PX-2127

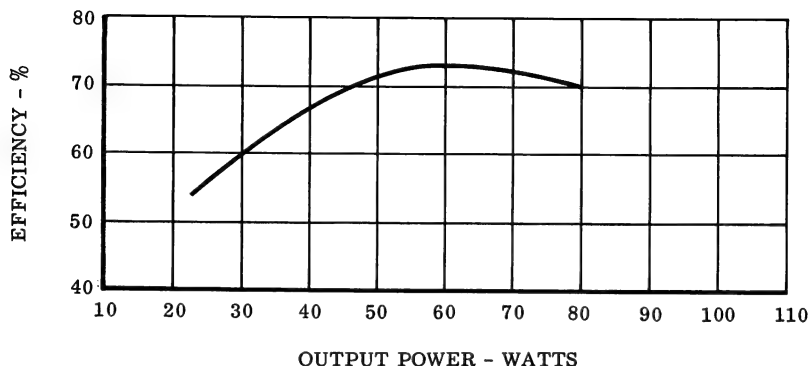
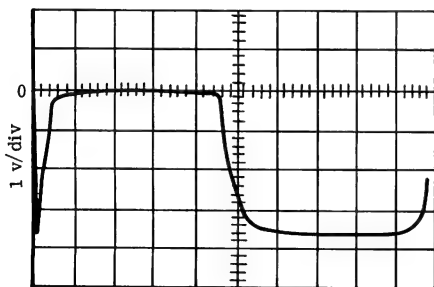


Figure 7-1-4 Efficiency Versus Power Output

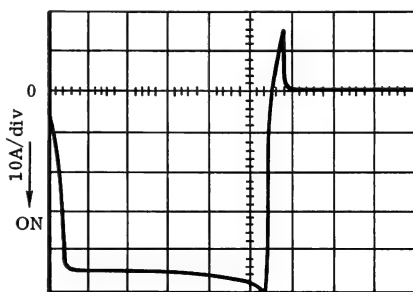
The collector-to-emitter voltage of this circuit would swing 4 volts peak-to-peak if there were no circuit losses. However, wiring resistances and saturation voltage of the transistor result in a voltage drop of about 0.4 volt when the circuit is switching 50 amperes. This can be seen from Figure 7-1-5. The saturation voltage is 0.1 volt and 0.3 volt is dropped due to wiring resistance.

The collector current of this circuit would alternate from zero to some peak value determined by the supply voltage and the impedance reflected back to the primary by the transformer. Figure 7-1-6 shows the collector current when switching 50 amperes peak. The current rise time is approximately 30 microseconds and fall time is about 20 microseconds. The slight hook at the 50 ampere peak is due to the characteristics of the core material of the transformer. The current spike in the opposite direction of the collector current is caused by inductance of the current sensing resistor. A 50-ampere meter shunt was used in obtaining this waveform.



.1 m sec/div

Figure 7-1-5 Collector-Emitter Voltage



.1 m sec/div

Figure 7-1-6 Collector Current

The base-to-emitter junction of the transistor is driven "on" to $V_{BE(sat)}$ of the transistor and back biased to 1.75 volts when driven "off". This waveform is shown in Figure 7-1-1. No protection is required across the base-to-emitter junction of the 2N2728, provided the back-bias is kept below 15 volts.

The base current was 2.5 amperes; therefore, the transistors are driven with a forced gain of 20 at 50 amperes. At currents above 50 amperes, the transistors do not fully saturate, which in turn causes a loss in efficiency.

An important check in inverter or converter design is the area the load line traverses. The load line shown in Figure 7-1-8, was obtained at 50 amperes. The safe area of operation of the 2N2728 is rectangular from 5 volts to 50 amperes for switching applications of this type. As can be seen, the transistor operates within the safe area which insures reliable operation.

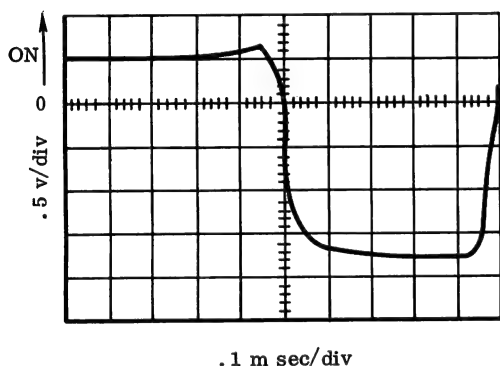


Figure 7-1-7 Base-Emitter Voltage

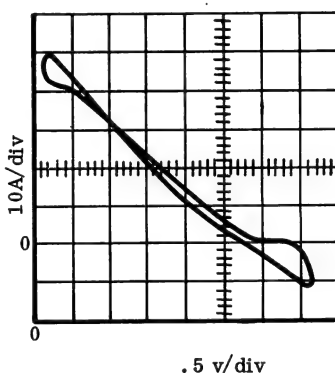


Figure 7-1-8 Load Line

2N2912 INVERTER AND CONVERTER CIRCUITS

The Motorola 2N2912 is another transistor suited for inverter and converter service. The characteristics of the 2N2912 transistor are such that it has low saturation voltage and is a fast switch. The transistor is packaged in a TO-8 case which enhances package size.

A PORTABLE ULTRA-VIOLET LIGHT SOURCE

The converter circuit shown in Figure 7-1-9 was designed to supply 400 ma at 26 volts from a 2.5 volt rechargeable battery. The 2N2912 transistor is saturating at less than .1 volt with 4 amperes of collector current and has a switching off speed less than .8 microseconds. This means that the power dissipation of the transistor with 8 kc operating fre-

quency is approximately 600 milliwatts and no heat sink is necessary. The transistor thermal resistance (junction to ambient) is approximately $30^{\circ}\text{C}/\text{watt}$ which will result in about an 18°C rise in junction temperature above ambient.

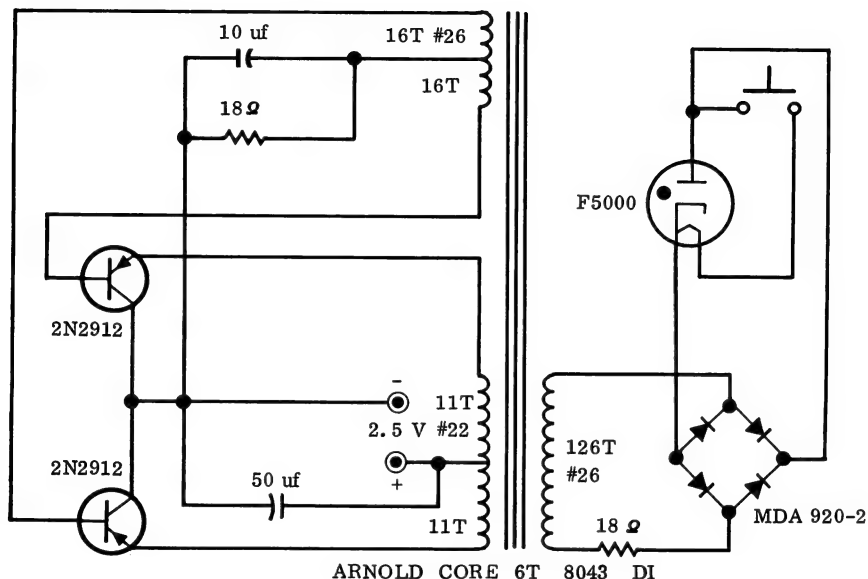


Figure 7-1-9 A Portable Ultra Violet Light Source

The drive circuit configuration gives excellent starting characteristics at all practical operating temperatures. The relatively high voltage source of the drive circuit results in rapid switching characteristics for the transistors. Because the drive source is referenced to the collector of the transistor, off drive voltage depends upon the additional voltage developed across the parallel RC circuit to assure turn-off. The total voltage developed between base and collector must be greater than twice battery voltage in order to reverse bias the base-emitter junction of the transistor. At low battery input voltage levels and with high gain transistors the power loss in the base drive resistor is small. At higher input voltage levels and with lower gain transistors the power loss in this base resistance can impair the operating efficiency.

Because of the rapid switching characteristics of the 2N2912, the optimum frequency depends mostly upon the core loss and the rectification efficiency. Considerations of core size and core loss resulted in the choice of a Deltamax material, and an 8 kc frequency. The Motorola MDA 920-2 bridge rectifier will cause some reduction in rectification efficiency at this frequency but the effect upon the entire circuit efficiency will be small.

A FOUR-WATT INVERTER CIRCUIT

An inverter circuit using the same basic characteristics as the converter circuit is shown in Figure 7-1-10. The circuit will supply 4 watts at 150 volts. The efficiency of the circuit is 80 per cent.

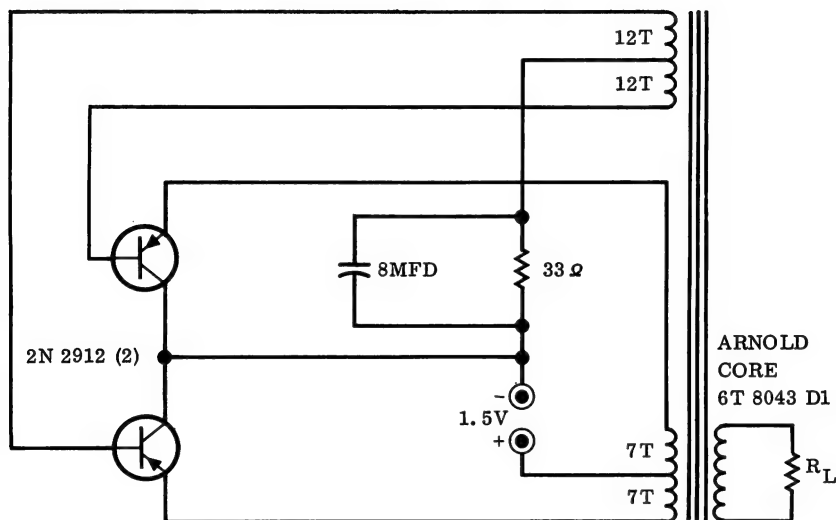


Figure 7-1-10 A 4 Watt 8KC Inverter

SECTION 7-2

High Frequency Inverter

INTRODUCTION

The circuit discussed here converts 28 volts dc to a 15 kc square wave. The output voltage can be set at the desired voltage level by using the number of secondary turns required for the specific output voltage.

The fast switching capabilities of the 2N2833 transistors enable efficient switching at 15 kc which permits the weight and size of reactive components to be reduced.

CIRCUIT DESCRIPTION

The circuit is a one transformer inverter operating in the common-emitter configuration as shown in Figure 7-2-1.

Both transistors were mounted on a single heat sink with the collectors uninsulated. The heat sink used has thermal characteristics similar to the MS-10 heat sink ($3^{\circ}\text{C}/\text{watt}$).

Resistors R_1 and R_3 as well as R_2 and R_4 form voltage dividers which provides 0.35 volts of base voltage to start the circuit. Resistors R_1 and R_2 limits the base current once the circuit begins to oscillate.

Capacitors C_1 and C_2 were used to speed-up turn on and off of the transistors to minimize switching times.

Diodes D_1 and D_2 are emitter-base clamping diodes. These diodes were necessary to keep from reverse biasing the emitter-base junctions beyond the rating of the 2N2833 transistors. In most inverters using 2N2833 transistors; this protection will be required since the emitter-base junction cannot be reverse biased in excess of two volts.

Zener diodes D_3 and D_4 are necessary to keep the collector-to-emitter voltages of the transistors at a safe level. The fast switching speeds cause voltage spikes due to leakage inductance in the circuit. These generated voltage spikes can break down the transistor if protective zener diodes are not used.

In using the diodes D_1 through D_4 , it was found that these components must be mounted physically as close as possible to the transistors. If this is not done, the lead inductance defeats the purpose of using the devices for protection.

The capacitor C_3 was used to eliminate lead inductance of the power supply leads. Unless this capacitor was used, the circuit would not function properly.

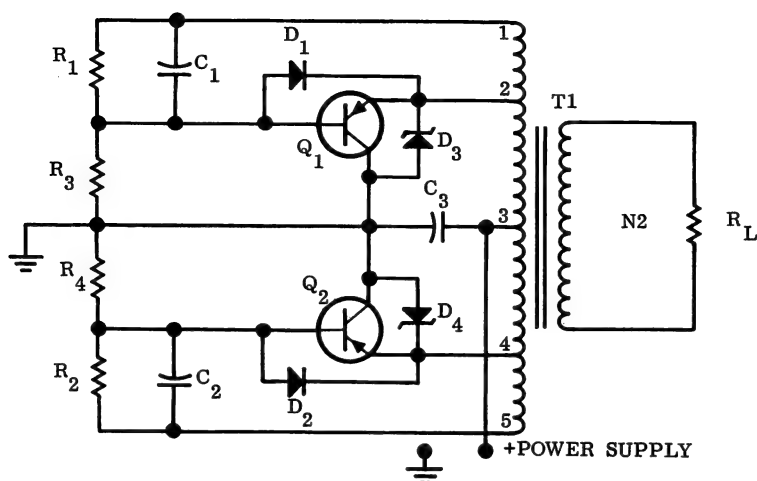


Figure 7-2-1 High Frequency Inverter

TABLE 7-2-1

Parts List for Figure 7-2-1

R_1, R_2	27 Ω , 1 W
R_3, R_4	2 2 K Ω , 2 W
C_1, C_2	0.5 μ f at 10 V
C_3	250 μ f at 50 V
D_1, D_2	1N4001
D_3, D_4	1N3039 B
Q_1, Q_2	2N2833
T_1	See Figures 7-2-2 and 7-2-3
Power Supply	28 V dc at 5 Amperes for 100 watts out

The frequency of oscillation of the circuit is determined by the transformer design. The transformer was designed to oscillate at 15 kc. The frequency of oscillation can be calculated from:

$$f = \frac{V_{CC} \times 10^8}{4 N_1 A_e B_s} \quad (7-2-1)$$

N_1 = the number of turns on one-half of the primary.

A_e = the effective cross sectional area of the core in square centimeters, (product of the physical area and the stacking factor).

B_s = the saturating flux density of the core in lines per square centimeters.

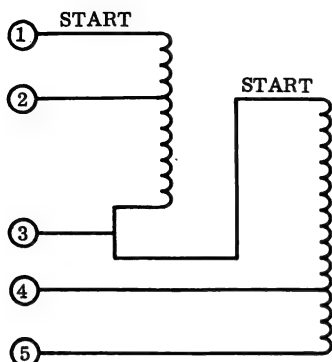


Figure 7-2-2 Transformer Winding Connections

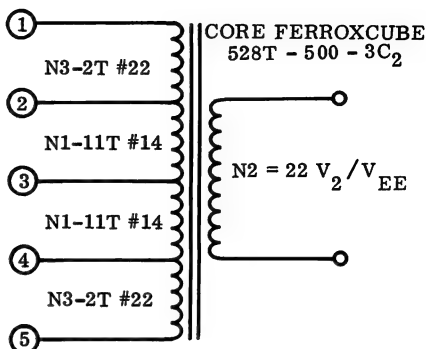


Figure 7-2-3 Inverter Transformer

A ferrite core was used to minimize eddy current and hysteresis losses which are a function of frequency. The core used was a Ferroxcube 528T-500-3C2 toroid. The primary consisted of 11 turns of #14 wire and the feedback winding was two turns. The windings were bifilar wound to minimize leakage inductance. Schematic representations of the transformer are shown in Figures 7-2-2 and 7-2-3.

The efficiency versus power output of the inverter is shown in Figure 7-2-4. At power levels of 100 watts, the circuit operates at an efficiency of 85 per cent. At power levels above 120 watts, the efficiency begins to decrease. The reason is because the transistors are not fully saturated at these power levels. At power levels below 100 watts, the efficiency also decreases. This decrease is due to overdriving the transistors. At power levels below 80 watts, the base current should be reduced for maximum efficiency.

The emitter-to-emitter voltage as a function of output power is also shown in Figure 7-2-4. As the power output is increased, the emitter-to-emitter voltage decreases because the transistors are not driven fully into saturation. From this figure, the number of secondary windings can be determined for a specified output voltage at a given power level by using the following equation:

$$N_2 = \frac{22 V_2}{V_{EE}} \quad (7-2-2)$$

Where

V_2 = the specified RMS output voltage

N_2 = the number of secondary turns required to obtain an output voltage of V_2 .

V_{EE} = the RMS emitter-to-emitter voltage at a particular power level determined from Figure 7-2-4.

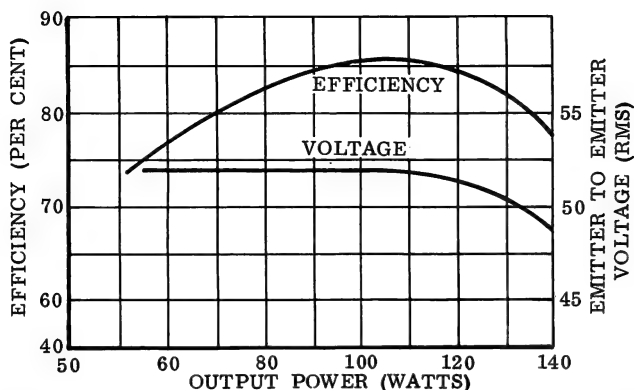


Figure 7-2-4 Efficiency and Output Voltage Versus Power Output

TRANSISTOR PERFORMANCE

The turn-off waveform of the collector-to-emitter voltage is shown in Figure 7-2-5. The effect of the zener diode across the collector-to-emitter of the transistor can be seen. The zener diode clamps the collector-to-emitter voltage at 62 volts for 0.5 microseconds.

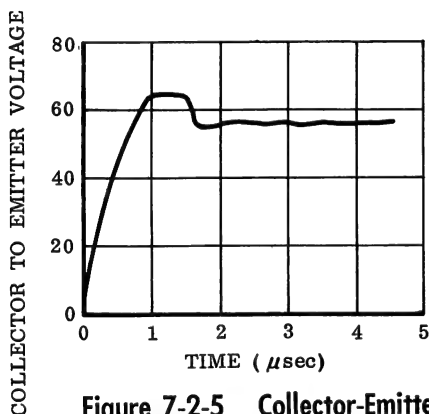


Figure 7-2-5 Collector-Emitter Voltage Versus Time

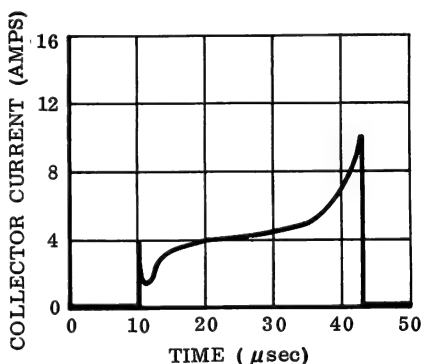


Figure 7-2-6 Collector Current Versus Time

The collector current the transistor switches is shown in Figure 7-2-6. The slight peak that occurs during turn on is due to larger amounts of base current during the charge time of the RC network in the base circuits. The spike that occurs just before turn-off of the collector-current when the transformer saturates and the voltage fed back, begins to decrease. When this occurs, the base current to the transistor decreases and the transistor begins to turn off. The peak collector current during the time the transformer saturates is limited only by the gain of the transistor. The peak current is $h_{FE}I_B$.

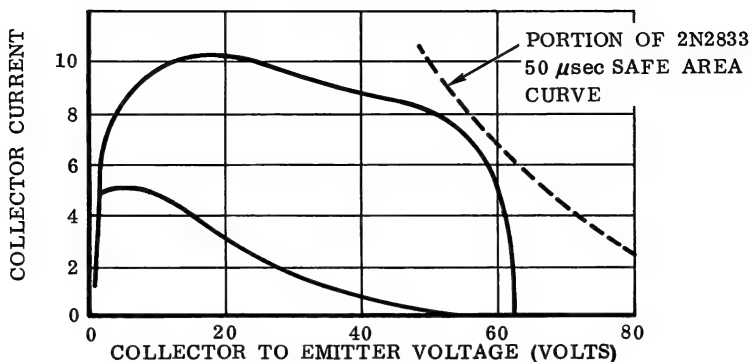


Figure 7-2-7 Load Line

The load line of the transistor is shown in Figure 7-2-7 with the circuit loaded to 100 watts. The load line is within the 50-microsecond safe operating area of the 2N2833 transistor. This will insure reliable performance for the 2N2833 transistor at 100-watt loads. For higher output power or reactive loads, the load line should be checked and a proper transistor selected.

SECTION 7-3

A Blocking Oscillator Fuel Pump Drive

The majority of electric fuel pumps use breaker contacts or commutating brushes to perform the switching function. In this section, a blocking oscillator circuit used to drive an electric fuel pump is discussed.

CIRCUIT DESCRIPTION

The circuit shown in Figure 7-3-1 utilizes a PNP power transistor in a blocking oscillator circuit to drive a solenoid plunger type fuel pump. The solenoid plunger assembly is a commercially available Bendix electric fuel pump. The solenoid drive was revised to include a feedback winding to complete the transistor oscillator circuit.

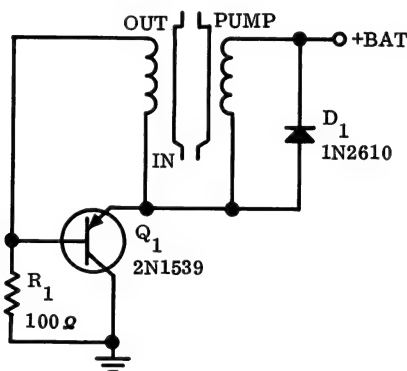


Figure 7-3-1 Basic Fuel Pump Oscillator Circuit

The circuit operation may be explained by first assuming that the transistor Q_1 has been driven into saturation by the feedback winding. The increasing collector current eventually reaches a value which can no longer be sustained by the decreasing base drive caused by the increasing saturation of the magnetic flux path. At this point, the feedback voltage begins to change polarity, and the collector current begins to turn off. The induced voltage in the feedback winding reverses its polarity assuring turn off of the transistor. The silicon diode D_1 is used to clamp the voltage of the transistor to approximately one volt above the battery level while dissipating the energy stored in the solenoid coil. When the energy is dissipated, the cycle of operation repeats. The 100 ohm resistor, R_1 , assures enough transistor current to start the next cycle.

The solenoid plunger assembly is drawn into the coil when the transistor is on and the hollow core of the plunger is filled with fuel. When the transistor turns off, a spring pushes the fuel filled plunger out of the coil

a predetermined distance. Thus, the plunger makes one excursion for each electrical cycle. The spring pressure exerts a pressure of approximately six pounds per square inch at the outlet port of the fuel pump. The valving in the plunger body allows fuel flow in one direction only.

CIRCUIT PERFORMANCE

It was determined that the transistor had to remain on for about 25 milliseconds to draw the solenoid the full distance back into the coil. This restricts the maximum frequency of operation. With the diode clamp across the solenoid coil, the L/R time constant is quite large and the coil current flows for a long period of time. The L/R time constant is reduced by adding a 10 ohm resistor R_2 in series with the diode as shown in Figure 7-3-2. This decreases the off time which, increases the frequency of operation, thus, the pumping rate is increased.

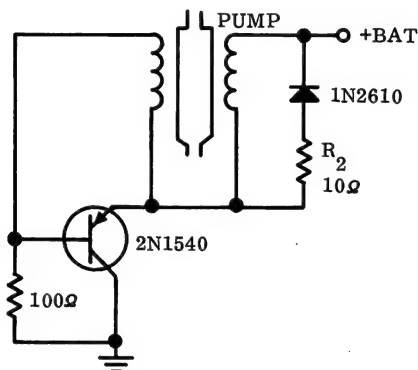


Figure 7-3-2 Fuel Pump Oscillator Circuit with Diode Resistor Network

A 30-volt zener diode connected across the transistor as shown in Figure 7-3-3 will reduce the coil current even faster, as well as protect the transistor against harmful voltage levels, however, both the diode-resistor and zener clamp necessitate the use of a slightly higher voltage transistor to assure reliable performance.

The fuel pumping rate with just the diode clamp was 11.8 gallons per hour. With the zener diode, the pumping rate was increased to 15 gallons per hour. This pumping rate was based on lifting the fuel 18 inches through five feet of hose.

The frequency of operation was designed for 16 cycles per second with a battery voltage of 13.5 volts. There is a slight dependence of frequency of operation on the supply voltage, however, the frequency deviations are small for the variations of a normal automobile battery.

LOAD LINE ANALYSIS

Shown in Figure 7-3-4 is the load line of the transistor operating in the circuit of Figure 7-3-3. The load line is the main consideration in selecting a transistor with a proper voltage rating. For the coil used, a 2N1540 transistor was the most economical choice. The load line is within the safe operating area of the 2N1540 transistor, therefore, reliable operation is assured.

When only the diode clamp is used, the voltage across the transistor is restricted to one volt above the battery voltage. In this case, a lower voltage transistor, similar to a 2N1539 can be used.

When a resistor is placed in series with the diode to increase the operating frequency, the load line will have to be checked, and a transistor with a proper voltage rating used.

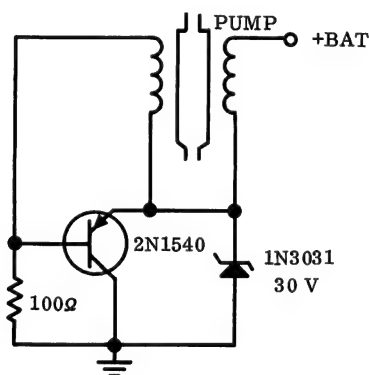


Figure 7-3-3 Fuel Pump Oscillator Circuit with Zener Diode

TEMPERATURE EFFECTS

In testing the circuit at low temperatures, it was found that pump starting was marginal which indicates that the number of feedback turns should be increased. It was found that a ratio of primary to feedback turns of 4 to 1 was necessary to insure proper starting at the lowest vehicle operating temperatures. A resistance in series with the feedback winding can set the proper on drive for the solenoid coil.

Increasing the number of feedback turns also increases the fuel pumping rate slightly by decreasing the off time of the blocking oscillator.

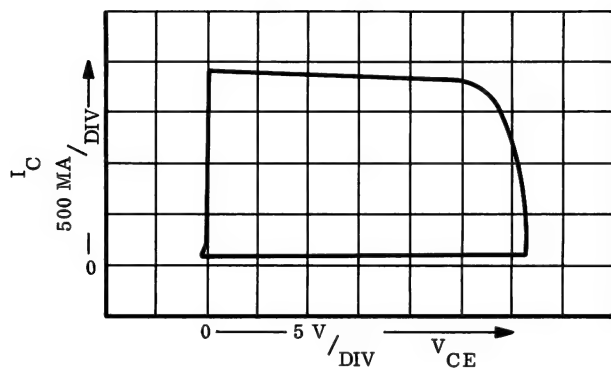


Figure 7-3-4 Transistor Load Line with Zener Diode Circuit

SECTION 7-4

SCR Inverters

The SCR has a number of characteristics which make it a desirable device for inverter applications. It has high voltage and current capabilities, and is not limited by a restricted "safe area" as is a transistor; that is, it can switch from maximum blocking voltage to maximum on current with any type of load. As with other devices, SCR inverters may take many forms. Since the SCR resembles the thyatron tube in its operation, SCR inverters have the same general form as thyatron inverters. However, since SCR's have lower conducting forward drop, higher current capabilities and faster switching times than thyatrons, the circuit constants will be somewhat different.

PARALLEL INVERTER

The parallel inverter most commonly makes use of capacitor commutation for turn-off. The basic principle of capacitor commutation is illustrated in Figure 7-4-1(a). The cycle is initiated by applying a trigger signal to the gate of the SCR, turning it on. This occurs at time t_1 in Figures 7-4-1(b) and (c). The voltage across the SCR drops to the forward conducting value in the order of 1 volt, and the anode current of the SCR rises to the on current as determined by the supply voltage V and the load resistor R_L . In addition, side A of the commutating capacitor C_C is brought to approximately ground potential, allowing it to charge through R_C with the polarity as indicated. It will charge to a voltage equal to the supply voltage minus the forward drop of the SCR. S_1 is now closed at the time designated as t_2 in the figures. The commutating capacitor C_C is connected across the SCR with polarity such as to reverse bias it. The SCR will momentarily conduct current in the reverse direction until its outer two junctions are swept free of carriers. This period is referred to as the reverse recovery time

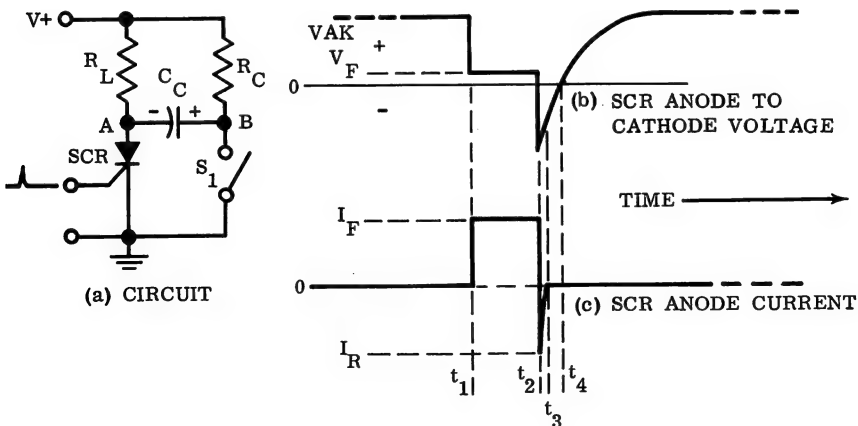


Figure 7-4-1 Capacitor Commutation

of the SCR and similar to that of a conventional diode. Note that the voltage across the SCR does not rise instantaneously to a value equal to the supply voltage. It is delayed by the charging of the commutating capacitor through R_L . This is a necessary delay since the SCR does not immediately regain its forward blocking capabilities upon the interruption of current or even upon the cessation of reverse current. For this reason $t_4 - t_2$ must be greater than or equal to t_0 , where t_0 is the turn off time of the SCR.

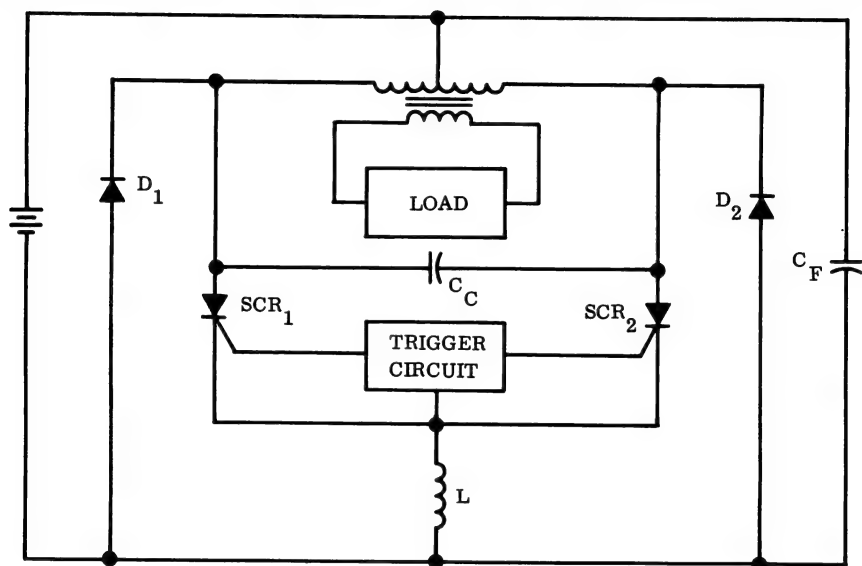


Figure 7-4-2 SCR Parallel Inverter

S_1 may be replaced by a second SCR. By the addition of an output transformer, the two SCR's may be made to share the load as well as serving to turn each other off. A circuit of this type appears in Figure 7-4-2. Note that in addition to the second SCR, several other components have been added. The inductance L aids in turn off since it opposes the instantaneous rise of load current during the commutation interval, as would be the case in Figure 7-4-1a. Consequently, the required value of the commutating capacitor is reduced. Diodes D_1 and D_2 prevent either end of the primary of the output transformer from rising above a voltage equal to twice the supply voltage by preventing the opposite end of the transformer from going negative. This condition would tend to occur during the switching interval due to the ringing of C_c and L , particularly under conditions of light or no load. The diodes, therefore, improve the operation of the inverter under lightly loaded conditions, and since they prevent the output voltage from rising as the load is decreased, they improve the regulation of the inverter. The filter capacitor C_F was added to maintain the supply voltage at the constant level during the switching interval.

The required value of commutating capacitor may be calculated from the equation

$$C_C (\text{min}) \approx 1.4 \frac{t_o I_A}{V}$$

where t_o is the turn off time of the SCR

I_A is the anode current inducted by the SCR, just before commutation and V is the voltage to which the commutating capacitor is charged; in this case twice the supply voltage.

L should be chosen to resonate with C_C at a frequency whose half period is equal to or greater than the turn-off time of the SCR.

RESONANT COMMUTATED INVERTER

An inverter which takes advantage of the current reversal in a tank circuit for commutation is shown in Figure 7-4-3, along with typical waveforms which would be encountered in such a circuit. The circuit operates as follows: The load capacitor C_L is charged to a voltage approximately equal to twice the supply voltage by the charging choke L_C . At this point, the current through the charging inductor attempts to reverse but is prevented from doing so by the disconnect diode D_D , which isolates C_L from the supply. At any desired time after the capacitor is charged, the SCR may be triggered on. This corresponds to time t_o on the waveform diagrams. The SCR turns on and the anode to cathode voltage drops to the forward conducting value V_F , as is indicated in Figure 7-4-3c. This effectively connects the primary of T_1 across C_L , and the two resonate as indicated by the capacitor voltage and anode current waveforms between t_o and t_2 , the current reverses but is only conducted momentarily by the SCR until its outer two junctions recover. As soon as the voltage on C_L drops below the supply voltage, current begins to flow in L_C . With the current in the load tank interrupted due to the reversal of the SCR bias, current and voltage waveforms are once again determined by L_C and C_L . Note that

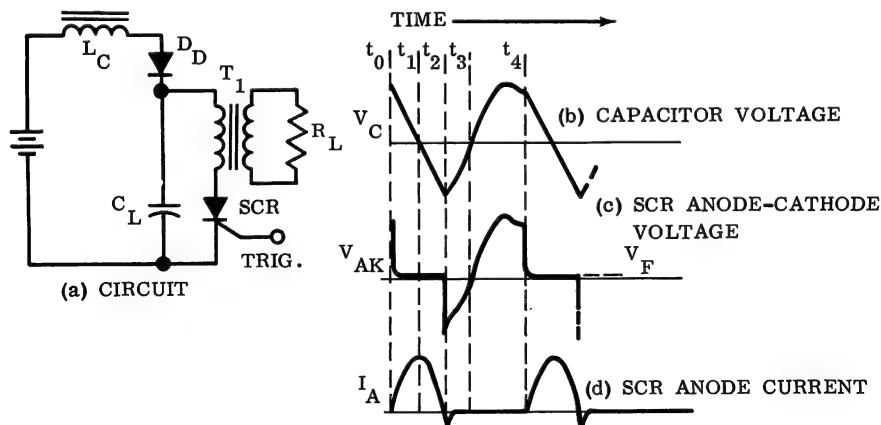


Figure 7-4-3 Resonant Commutated Inverter

the SCR remains reverse biased until t_3 . This delay is necessary to allow the SCR to regain its forward blocking capability. At t_4 the cycle is complete and repeats itself.

The circuit could operate similarly without D_D provided that triggering occurred at the proper time. A change in trigger time, however, would change the output voltage since the capacitor voltage is oscillatory. The inclusion of diode D_D allows the repetition rate to be reduced without changing the peak output voltage or current. In order to insure current reversal in the output tank circuit and consequent turn-off of the SCR, the tank circuit Q must be kept high, or to put it another way, the ratio of circulating current to reflected load current must be high. This may be achieved by using high Q components, by limiting the load current and by providing a small amount of leakage inductance. Since the current through L_C begins to build up as soon as the SCR is turned on, and this current tends to oppose the current reversal in the tank circuit, the resonant period of the charging circuit must be considerably longer than that of the output tank.

Since this circuit uses only one SCR and has fewer total components than the parallel inverter, as well as having simpler trigger requirements, its cost is considerably lower. This type of inverter is suitable for applications such as high voltage, low current power supplies, fluorescent light power supplies and ultrasonic generators.



SECTION 8

Power Supplies and Regulators

SECTION 8-1

Switching Regulators

Regulators can be classified into two basic types. The first type of regulator is the conventional type which controls the amount of direct current. The other type of regulator is a switching type regulator, which will be discussed in this section.

The prime reason for utilizing a switching regulator is to minimize power losses in the series pass transistors which occur when large variations of input or output voltages are encountered.

THEORY OF OPERATION

The switching regulator depends on the fact that the output voltage is directly proportional to the input voltage by the duty cycle of the switched voltage. This relationship can be established by considering the voltage waveform of Figure 8-1-1 applied to the network in Figure 8-1-2.

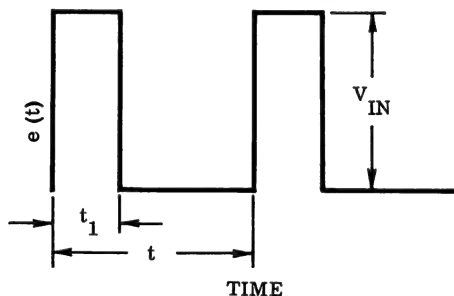


Figure 8-1-1 Non-Symmetrical Driving Voltage

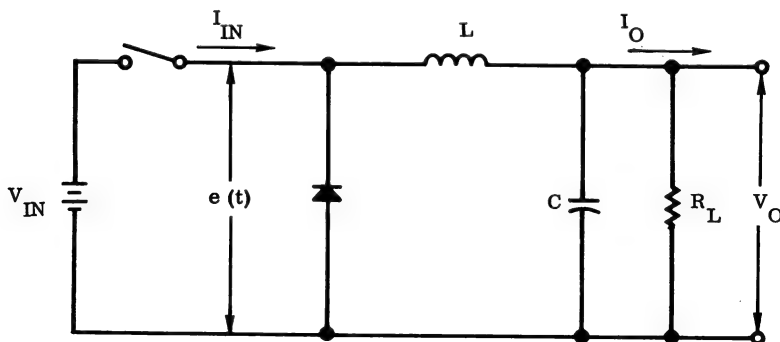


Figure 8-1-2 Basic Switching Regulator Circuit

Assuming that the inductor is of sufficient value to keep the change in input current relatively small during the on time of the voltage pulse, the average input power will be:

$$P_{in} = V_{in} I_{in} t_1/t \quad 8-1-1$$

$$I_{in} = I_o \quad 8-1-2$$

where I_{in} is the peak input current.

If the filter capacitor is of sufficient magnitude to keep the ripple small during the off time, the output power will be:

$$P_o = V_o I_o \quad 8-1-3$$

Assuming no losses during the switching interval, the average input and output powers are equal. Combining equation 1, 2, and 3, the following relationship is established:

$$V_o = V_{in} (t_1/t) \quad 8-1-4$$

The value (t_1/t) is the duty cycle of the input pulse.

CIRCUIT DESCRIPTION

The regulator is shown in block diagram form in Figure 8-1-3 to simplify the circuit description. Shown in Figure 8-1-4 is the complete schematic. The associated parts list of the regulator is shown in Table 8-1-1.

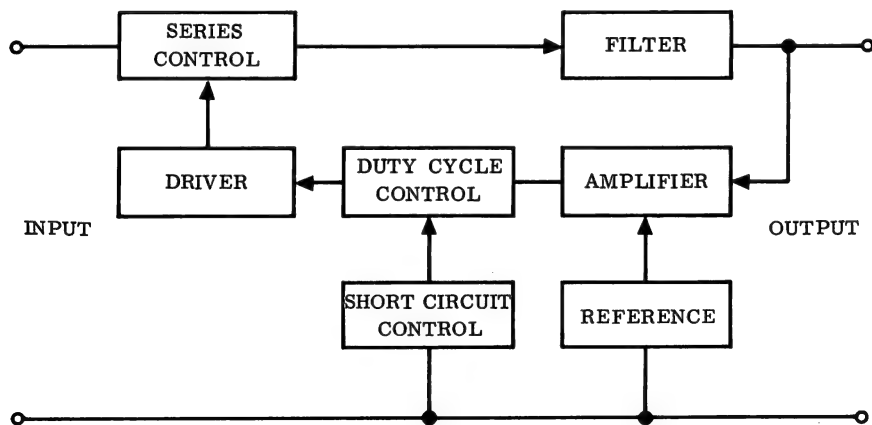


Figure 8-1-3 Switching Regulator Block Diagram

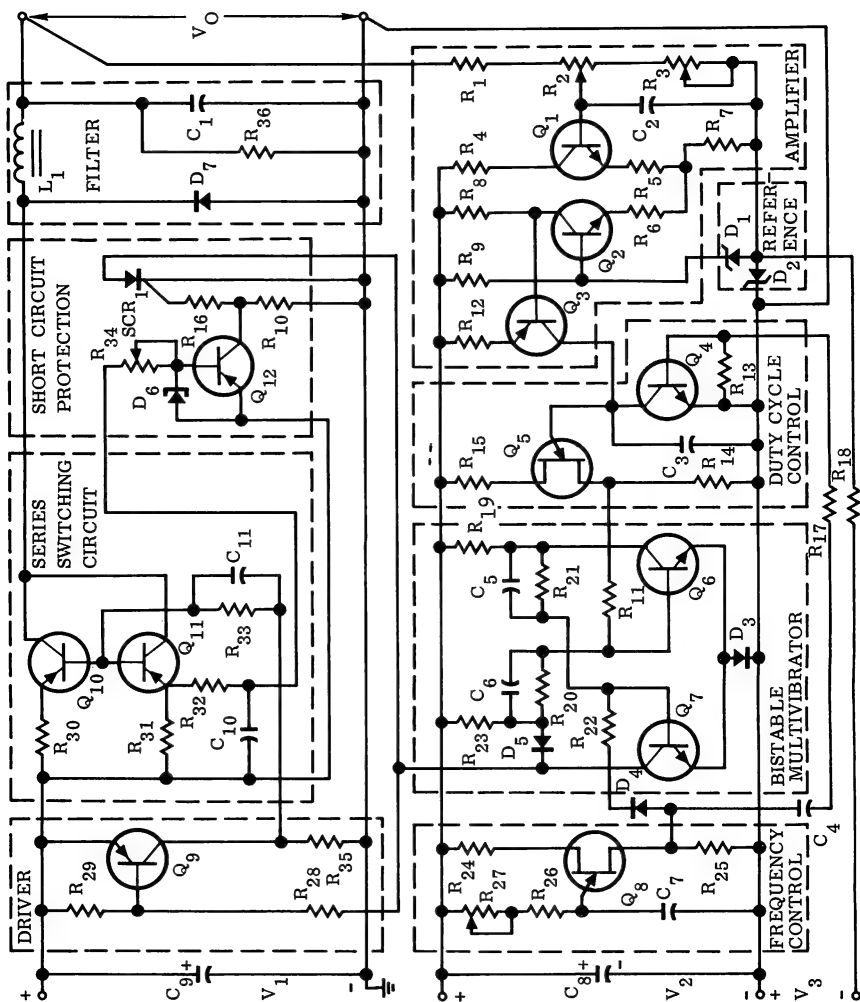


Figure 8-1-4 Switching Regulator

TABLE 8-1-1
PARTS LIST FOR SWITCHING REGULATOR

<u>Resistors and Potentiometers</u>			<u>Capacitors</u>		
R ₁	120	Deposited carbon ± 1%	C ₁	12, 500	at 75V
R ₂	500	10T Potentiometer	C ₂	0. 001	
R ₃	5	Potentiometer	C ₃	0. 003	
R ₄	2. 2K	Deposited carbon ± 1%	C ₄	0. 05	
R ₅ , R ₆	100	Deposited carbon ± 1%	C ₅	2200pF	
R ₇	12K	Deposited carbon ± 1%	C ₆	. 01	
R ₈	10K	Deposited carbon ± 1%	C ₇	0. 006	
R ₉	2. 7K	Deposited carbon ± 1%	C ₈	50	at 25V
R ₁₀ , R ₁₃ , R ₁₆ , R ₁₈	4. 7K		C ₉	2000	at 75V
R ₁₁ , R ₁₄ , R ₂₅	100		C ₁₀	0. 1	
R ₁₂	150	Deposited carbon ± 1%	C ₁₁	50	at 15V
R ₁₅ , R ₂₄	330		All capacitors are ± 10%. The value of capacitance is micro-farads unless otherwise specified.		
R ₁₇	5. 6K		<u>Semiconductors</u>		
R ₁₈ , R ₂₁	1. 8K		D ₁ , D ₂	1N823A	
R ₁₉ , R ₂₀ , R ₂₃	1K		D ₃ , D ₄ , D ₅	1N4001	
R ₂₂	470		D ₆	1N3128	
R ₂₆	36K		D ₇	1N3889	
R ₂₇	10K	Carbon potentiometer			
R ₂₈	330	5W Carbon	Q ₁ , Q ₂ , Q ₄ , Q ₆ , Q ₇	2N697	
R ₂₉	20		Q ₃	2N2800	
R ₃₀ , R ₃₁	0. 1, 5W, W. W.		Q ₅ , Q ₈	MU970	
R ₃₂	47		Q ₉ , Q ₁₀ , Q ₁₁	2N2832	
R ₃₃	2, 5W, W. W.		Q ₁₂	2N2043	
R ₃₄	100, 10T Potentiometer		SCR-1	MCR 914-1	
R ₃₅	40, 50 Watt, W. W.		<u>Inductors</u>		
R ₃₆	100, 5W, W. W.		L ₁	8 mH at 10A dc Triad C-49U	
All resistors are 1/2 watt, ± 5% carbon resistors. The values are in ohms unless otherwise specified.			<u>Supply Voltages</u>		
			V ₁	45V ± 5V at 10 amps peak with a 50% duty cycle	
			V ₂	20V ± 5% at 75 mA	
			V ₃	20V ± 5% at 25 mA	

The amplifier consists of a differential amplifier and an additional stage of current gain. Transistors Q₁ and Q₂ form the differential amplifier. Degeneration in the differential amplifier was allowed to minimize voltage gain drift caused by hFE variations with temperature. Transistor Q₃ gives additional gain as well as giving the proper phase shift to drive the unijunction circuit.

The differential amplifier senses a portion of the output of the regulator and compares this voltage to the reference. The difference in the voltage is amplified by the amplifier. This amplified signal is applied to the base of Q₃ and controls the charging rate of C₃.

The reference circuit consists of a temperature compensated zener diode D₁ biased to its test current. The zener diode provides a stable voltage source for a reference.

The zener diode D₂ is used to provide an offset voltage for the reference and the differential amplifier. This allows the output voltage to be variable from zero to 20 volts. Without this offsetting voltage, the minimum output voltage would be the reference voltage. The design should allow the output to actually reverse polarity by choosing D₂ to be slightly higher in voltage than D₁ and increasing the size of R₃ for zero adjustment.

The duty cycle control circuit consists of transistors Q₄ through Q₈. The unijunction transistors Q₅ and Q₈ are used in relaxation oscillator circuits. These oscillator circuits generate pulses which trigger a bistable multivibrator circuit. The relaxation oscillator utilizing Q₈ controls the basic operating frequency of the regulator at 5 kc. The operating frequency can be calculated from the following equation:

$$f = \frac{1}{(R_{26} + R_{27}) C_7 \ln\left(\frac{1}{1-\eta}\right)} \quad 8-1-5$$

The term η is the intrinsic standoff ratio of the unijunction transistor and can vary for different devices. The resistor R₂₇ can be adjusted for oscillation at 5 kc. The unijunction transistor Q₅ sets the "on-off" duty cycle of the basic 5 kc repetition rate generating pulses at time intervals depending on the rate of charge of C₃. The relationship for the rate of charging the capacitor C₃ is:

$$\frac{dv}{dt} = \frac{I_{C3}}{C_3} \quad 8-1-6$$

I_{C3} is the collector current of Q₃.

Transistor Q₄ is used to discharge Capacitor C₃ at the beginning of each 5 kc trigger pulse. This resets the charging of the capacitor from zero at the beginning of each cycle. This enables the regulator to operate at light loads and have good load regulation. Transistors Q₆ and Q₇ make up

the bistable multivibrator. The pulses generated by the two unijunction circuits trigger the bistable multivibrator to generate variable duty cycle pulses. Diode D₃ in the emitters of Q₆ and Q₇ provides off bias for the transistors.

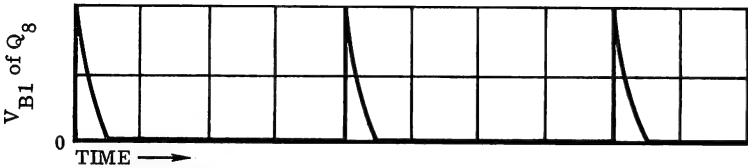


Figure 8-1-5A Q₈, R₂₅ = 5 kc Trigger Voltage

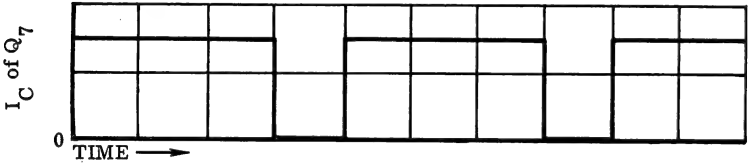


Figure 8-1-5B Q₇ - Collector Current

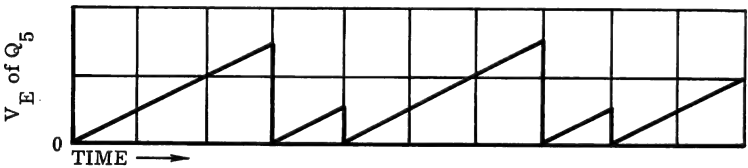


Figure 8-1-5C C₃ Voltage

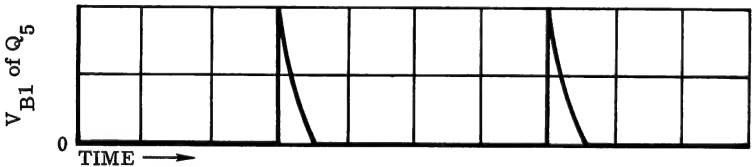


Figure 8-1-5D Q₅, R₁₄ - Control Trigger Voltage

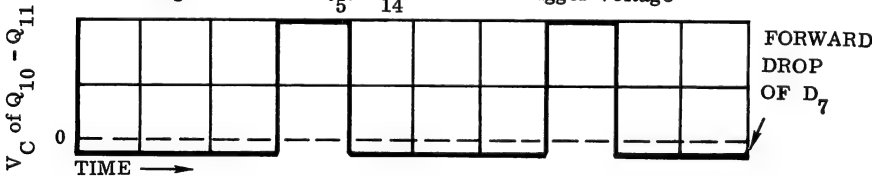


Figure 8-1-5E Applied Voltage to Filter

Figure 8-1-5 Circuit Waveshapes

(All Voltage are with Respect to Ground)

The driver circuit consists of a power transistor Q₉. This transistor is driven into saturation by the collector current of Q₇. When Q₉ is saturated, the series switching transistors are biased off due to the charge

across C₁₁. When the driver transistor is off, one ampere of base drive current is available to the series switching transistors.

The series switching transistors utilize a pair of power transistors operating in parallel. These transistors are biased on through R₃₅ and turned off by saturating the driver transistor. The RC network consisting of R₃₃ and C₁₁, is a speed-up network to minimize the switching time of the transistor as well as provide off voltage. Resistors R₃₀ and R₃₁ were used to provide current equalization in the series switching transistors operating in parallel.

To obtain an understanding of the operation of the switching regulator, the generation of one cycle of voltage into the filter circuit will be established with the aid of Figures 8-1-5A through 8-1-5E. The pulses shown in Figure 8-1-5A are the trigger pulses generated by the 5 kc relaxation oscillator, Q₈. These pulses trigger on Q₇ of the bistable multivibrator. The collector current of Q₇ is shown in Figure 8-1-5B. At the beginning of the 5 kc trigger pulse, the capacitor C₃ begins to charge from a current source as shown in Figure 8-1-5C. When the voltage across the capacitor approaches the firing voltage of the unijunction transistor Q₅, the capacitor is discharged through R₁₄ generating a pulse as shown in Figure 8-1-5D. This pulse triggers the opposite side of the multivibrator and turns Q₇ off until another pulse is applied from the 5 kc relaxation oscillator. This completes one cycle of operation of the duty cycle control circuit. The voltage that is applied to the filter circuit is shown in 8-1-5E. This voltage is averaged by the filter and diode to give a dc level out.

CIRCUIT PERFORMANCE

The basic function of a regulator is to hold the output voltage constant regardless of load, line and temperature variations.

The load regulation of the switching regulator is shown in 8-1-6. The load regulation of the switching regulator from no load to full load was 0.05%

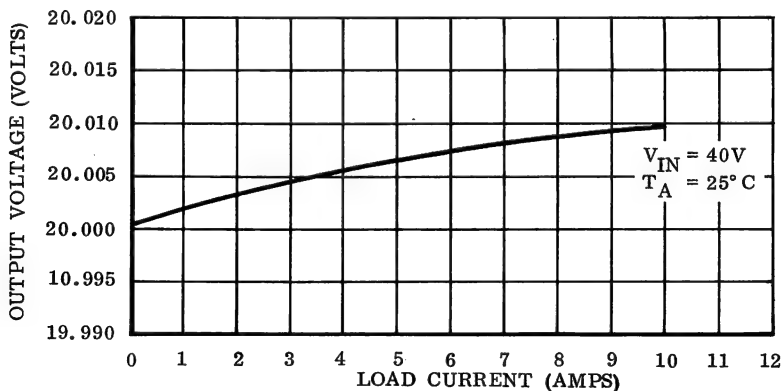


Figure 8-1-6 Load Regulation

The input regulation of the switching regulator is shown in Figure 8-1-7. With the circuit operating at 10 amperes and 20 volts out, the change in output was less than $\pm 0.07\%$ for a $\pm 10\%$ change in input voltage. If the input voltage drops below 40 volts, regulation is not as good.

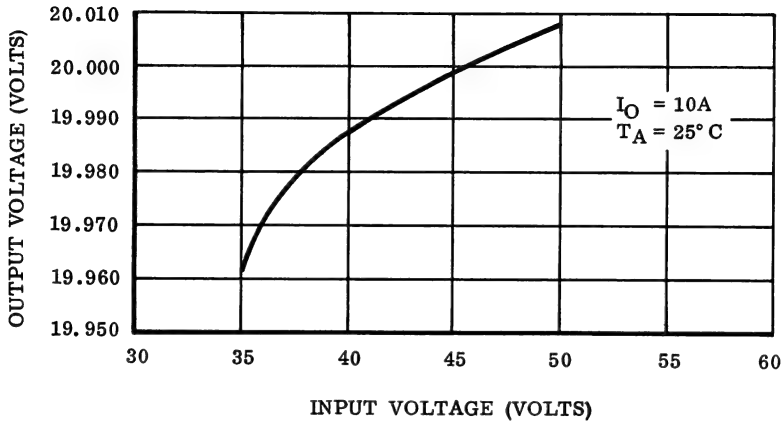


Figure 8-1-7 Input Regulation

Temperature effects on the regulator is shown in Figure 8-1-8 which is the plot of output voltage versus ambient temperature. In making the temperature measurements, it was found that the temperature coefficients of resistors in the sensing circuit will cause variation in the output. To minimize the effect of temperature on the output voltage, the resistors in the sensing circuit should have low temperature coefficients.

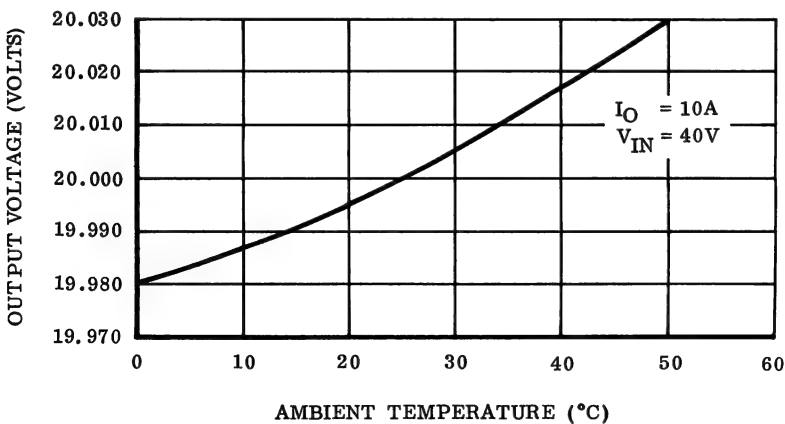


Figure 8-1-8 Temperature Regulation

The transient response of the regulator was in the order of 10 milliseconds. This response time is measured as the time required for the output voltage to recover within 10 per cent of the half load output level when loaded from half load to full load. This response time is dependent on the magnitude of the reactive components in the filter circuit.

When the output of this regulator circuit is unloaded from full load, a transient voltage appears when the load is removed immediately after the series switching transistor is turned on. The amplitude of this voltage is related to the reactive components by the following equation:

$$V_{o(max)} = V_1 + (V_o - V_1) \cos \frac{t_1}{\sqrt{LC}} + I_o \sqrt{\frac{L}{C}} \sin \frac{t_1}{\sqrt{LC}} \quad 8-1-7$$

V_1 is the maximum input voltage

V_o is the full load voltage

I_o is the full load current

t_1 is the maximum on time of series transistor.

For the values of L and C used in the circuit the maximum transient voltage was calculated to be 23.2 volts or 3.2 volts in excess of the normal 20 volts output. Actual transients were measured to be peaks of 2.8 volts above the normal out with full load to half load change in output current.

DESIGN CONSIDERATION

Power Dissipation in the Series Switching Transistor

The prime purpose for using a switching regulator is to minimize the power dissipation in the control transistor. The power dissipation in the switching series transistor will be determined in this section.

The idealized current and voltage waveforms of the series switching transistor are shown in Figure 8-1-9.

Switching of the voltage occurs before the current is switched because of the diode and inductor in the filter circuit. From the idealized waveforms, it can be shown that the collector dissipation of the series switching transistor can be approximated by equation 8-1-8.

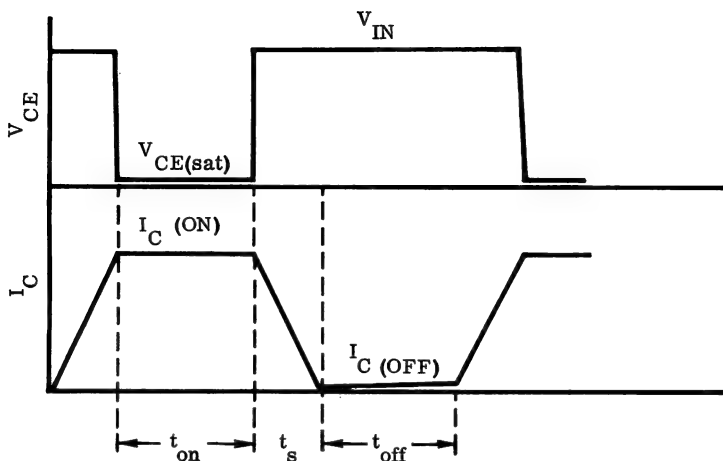


Figure 8-1-9 Idealized Waveforms of Series Switching Transistors

$$P_c = V_{CE(sat)} I_C T + V_{in} I_{C(off)} (1 - T) + V_{in} I_C t_s f. \quad 8-1-8$$

$V_{CE(sat)}$ is the saturation voltage of the transistor.

I_C is the peak collector current.

T is the duty cycle of the switched waveform; ratio of on time to period.

V_{in} is the dc input voltage.

$I_{C(off)}$ is the leakage current of the transistor during cutoff.

t_s is the switching time of the current waveform.

f is the operating frequency of the regulator.

The following are typical values of the circuit characteristics of equation 8-1-8 when the regulator is operating at an output of 20 volts and 10 amperes.

$$V_{CE(sat)} = 0.1 \text{ V}$$

$$V_{in} = 45 \text{ V}$$

$$I_C = 5 \text{ A}$$

$$f = 5 \text{ kc}$$

$$T = 45\%$$

$$t_s = 1 \mu\text{sec.}$$

$$I_{C(off)} = 10 \text{ mA}$$

The collector dissipation of each series switching transistor is approximately 2 watts. The switching speed of the 2N2832 transistors are fast enough for operation at 10 kc, however, the switching losses will double at 10 kc.

A conventional regulator operating at the same output power operating under ideal conditions will dissipate 15 to 20 watts, and if input voltage is high and load maximum, the power dissipation would be much more.

LOAD LINE OF SERIES SWITCHING TRANSISTORS

The load line excursion of the series switching transistor, under normal operating conditions is at its worst-case when operating at full load. The load line of one of the series switching transistors is shown in Figure 8-1-10.

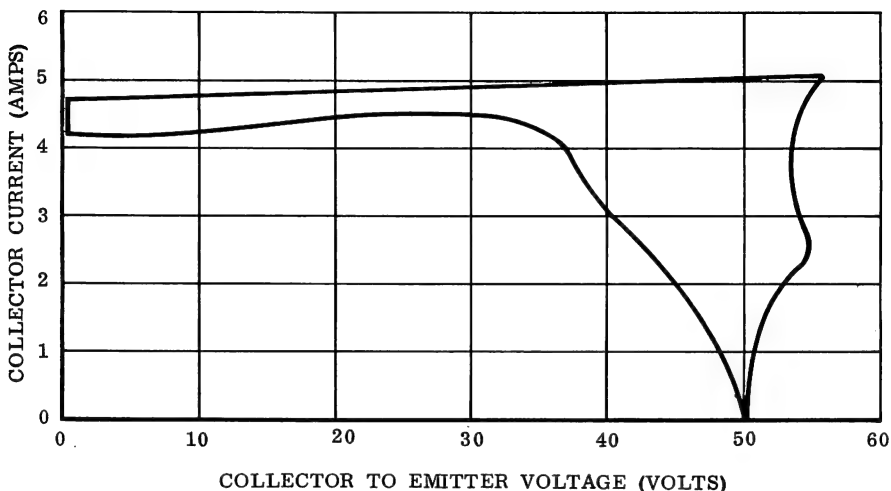


Figure 8-1-10 Load Line of the Series Switching Transistor

The worst load line excursion occurs at 5 amperes and 56 volts. This indicates that a pair of 2N2832 transistors will operate reliably in this circuit as the control transistors.

THE CURRENT CIRCULATING DIODE

The current circulating diode D_7 must have a fast recovery time. Unless this diode has a fast recovery time, high peak currents will be supplied by the transistor during the turn-on time of the transistor. This will cause high peak power excursions of the load line of the transistors. If this is the case, it may be necessary to parallel more transistors to restrict load line excursions within the safe area of the transistor, or to use a higher voltage device such as the 2N2834.

SHORT CIRCUIT PROTECTION

The short circuit protection incorporated in this regulator consists of a tunnel diode sensing circuit (See Section 8-2). The sensing circuit triggers a silicon controlled rectifier and shuts the series switching transistors off. Once the silicon controlled rectifier is fired due to a current overload, both the main supply input voltage (V_1) and the control supply (V_2) must be removed to reset the circuit.

SWITCHING SEQUENCE OF DC SUPPLY VOLTAGES

The regulator is designed such that the series switching transistor is normally saturated. Therefore, if V_1 is turned on before V_2 , the full voltage V_1 will appear at the output. To eliminate this problem, V_2 should be turned on first. In turning off the supply, the same problem occurs; therefore, V_1 should be turned off before V_2 .

The switching regulator has excellent static regulating characteristics, however, the transient response is limited by the reactive components in the filter circuit. The ultimate in a power supply can be achieved by using the switching regulator as a post regulator to a conventional direct current regulator. The switching regulator would be used to hold the voltage across the pass transistor to a constant value and minimize the power dissipation during line and load variations.

SECTION 8-2

SHORT CIRCUIT PROTECTION

The shortcoming of many regulated power supplies, using transistors as the control element, is that the transistors are not protected adequately in case of load malfunctions or accidental output short circuits. During short circuits, high peak currents and voltages are encountered by the control transistor. Unless this surge of power is limited or removed immediately, the transistors may operate beyond their limitations and become damaged. (See Section 5). This section deals with circuits that can be used to protect transistors in regulated power supplies against short circuited outputs.

THE CURRENT LIMITER

The simplest form of short circuit protection would be to have a resistor in series with the output of the power supply to limit the current to a reasonable level during a short circuit. The disadvantage of this technique is that at high current levels the dissipation of the resistor becomes intolerable.

To minimize the power loss during normal operation of a current limiter, a transistor can be used. A circuit of a current limiter is shown in Figure 8-2-1.

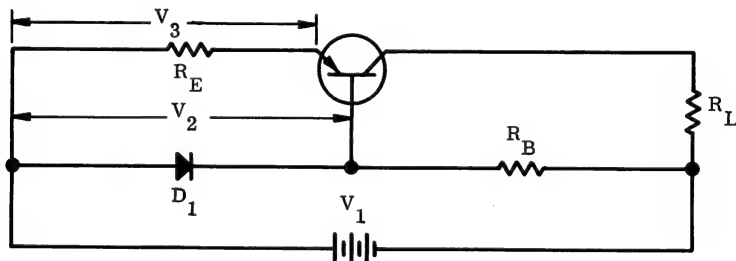


Figure 8-2-1 Current Limiter

During normal operation, the voltage $V_3 + V_{EB}$ is less than V_2 , therefore the diode, D_1 , is not in its conduction state. All the current through R_B can be assumed to be base current of the transistor. R_B is selected to insure saturation of the transistor. During this time, the power loss is the product of $V_{CE(sat)}$ and I_L . When the load is short circuited, the current increases but the diode clamps the voltage $V_3 + V_{EB}$ and the current is limited. The current at which the circuit limits can be calculated as

$$I_E = \frac{V_2 - V_{BE}}{R_E}$$

During the time the transistor is limiting the current, the output voltage is transferred to the transistor. The transistor can operate at high voltage and current, therefore, the transistor should have safe area characteristics capable of handling the load line excursions. The transistor should be mounted on an adequate heat sink to prevent excessive heating of the transistor during the short circuit.

The current limiter incorporated in a regulator circuit is shown in Figure 8-2-2.

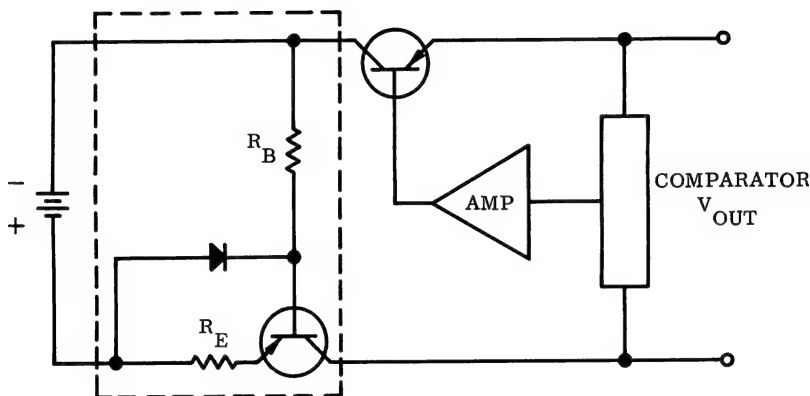


Figure 8-2-2 Power Supply With Current Limiter

THE ELECTRONIC FUSE

Another method of protecting transistors in power supplies is to open the circuit when a short circuit occurs. Circuit breakers and fuses are often used for this purpose, however, in the majority of cases these devices are not fast enough to protect transistors.

To overcome the long time delay associated with circuit breakers and fuses, a circuit that accomplishes the same function has been designed. The circuit is shown in Figure 8-2-3.

The circuit operation is as follows: When the supply voltage V_1 is turned on, Q_1 is saturated by the base current through R_3 . The load current causes a voltage drop across the sensing resistor R_1 . When the voltage across R_1 reaches approximately 0.8 volts, diode D_1 and transistor Q_2 begin to conduct. As conduction in Q_2 begins, the base of Q_1 is driven more positive causing the transistor to start to fall out of saturation. When the collector of Q_1 has gone sufficiently negative to allow D_2 to conduct, Q_2 will switch fully on and Q_1 will be fully off. The circuit is reset by removing the excess load and momentarily opening the reset switch. The circuit can also be reset by momentarily removing the power.

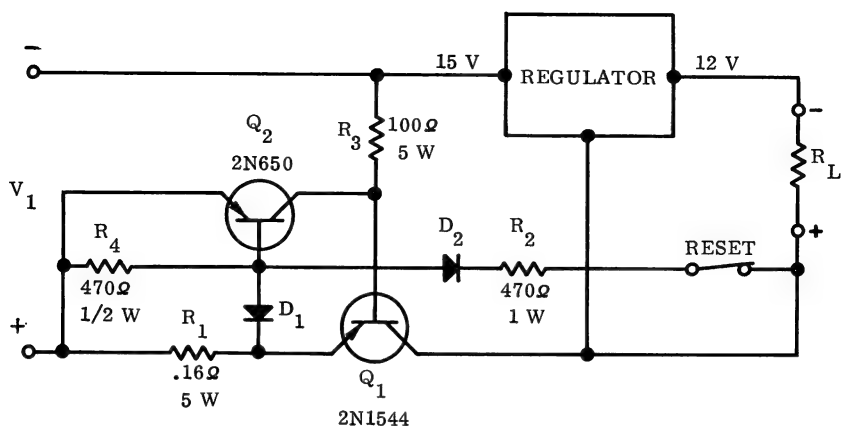


Figure 8-2-3 Electronic Fuse Circuit

The circuit shown in Figure 8-2-3 was intended to operate from a 7 to 12-volt 5-ampere regulated power supply. When the load current reaches 5.5 amperes, the transistor will open the circuit. The switching from on to off was accomplished in about 90 microseconds as indicated in Figure 8-2-4.

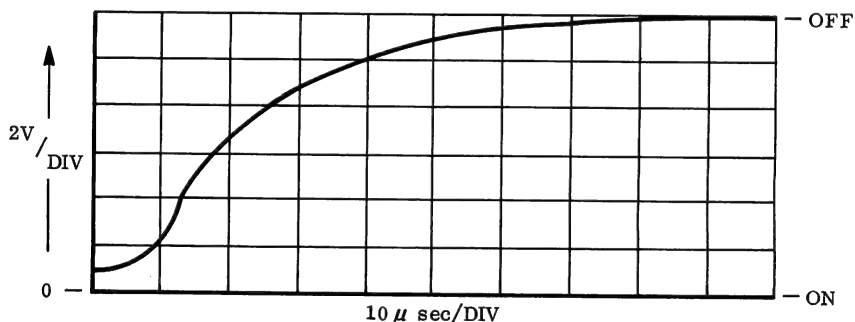


Figure 8-2-4 Turn Off Time of Electronic Fuse

The circuit offers fast protection with very little overload. There are no fuses or circuit breakers that are comparable to this circuit.

CURRENT SENSING WITH TUNNEL DIODE

One of the problems in sensing overload current in power supplies is to obtain a stable low level sensing circuit. The use of a tunnel diode in a threshold sensing circuit is one solution to this problem. The basic sensing circuit is shown in Figure 8-2-5.

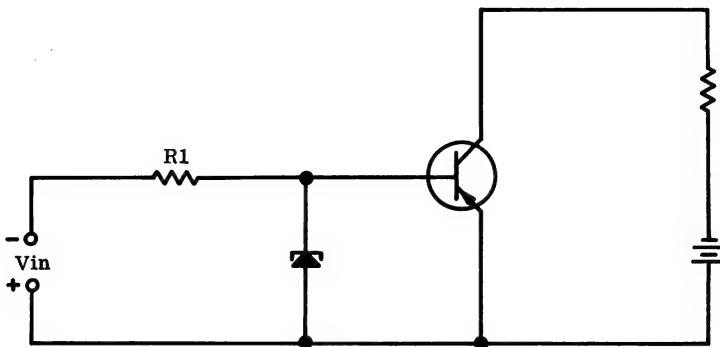


Figure 8-2-5 Tunnel Diode Sensing Circuit

The operation of the circuit can be understood by observing the V-I characteristics of the tunnel diode and the transistor that is shown in Figure 8-2-6 and 8-2-7. The voltage across the tunnel diode and the base-to-emitter of the transistor is the same; therefore, the composite characteristic is generated by addition of the device currents for a given voltage as shown in Figure 8-2-8. The resistor R_1 is selected such that the operating point Q_1 is below the peak point voltage and current for a given input voltage V_{in} . As V_{in} is increased, the operating point approaches the peak point. When the voltage is sufficient for the operating point to shift over the peak point, an operating point similar to Q_2 is established. The transistor has been switched from a cutoff state to a conducting state.

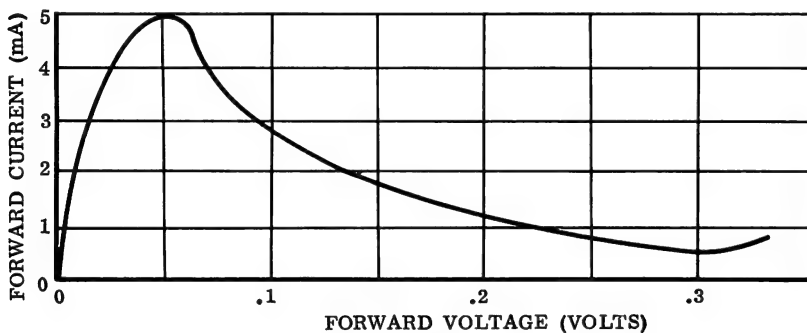


Figure 8-2-6 Tunnel Diode Characteristics

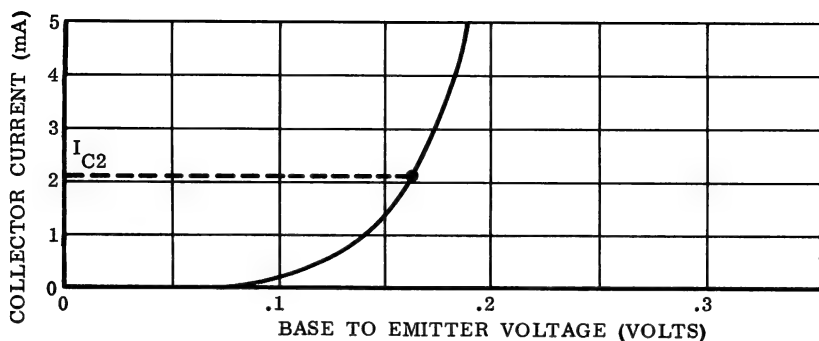


Figure 8-2-7 Transconductance Characteristics

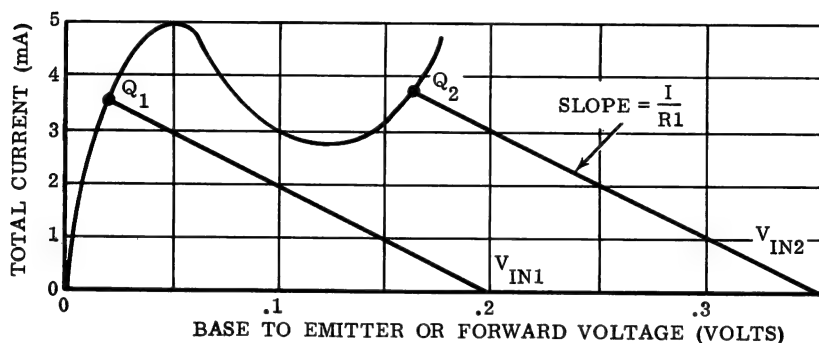


Figure 8-2-8 Composite Characteristic

The tunnel diode allows the transistor to switch from cutoff to a current of sufficient amplitude to control amplifiers in regulators or trigger higher power devices used for short circuit protection. For example, a series pass transistor could be quickly biased to cutoff.

This type of short circuit protection is used in the switching regulator discussed in Section 8-1, and the motor control in Section 6-4.

SCR CROWBAR CIRCUIT PROTECTION

The reaction time of fuses and circuit breakers is too slow to protect transistors and other solid state devices. In fact, if the faults are over voltage, circuit breakers and fuses may not draw sufficient current to open at all.

Using a zener diode with an SCR and a circuit breaker will clear over voltage faults. Both dc and ac circuits can be protected as shown in Figure 8-2-9.

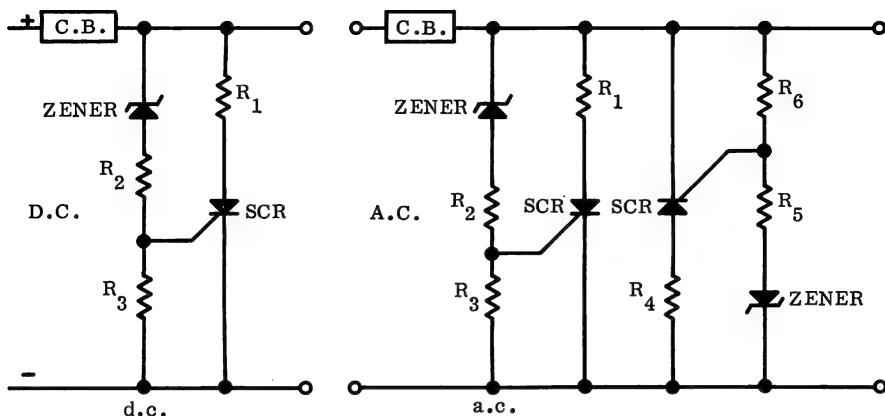


Figure 8-2-9 SCR Crow Bar Over-voltage Protection Circuit

The SCR surge current must be limited by R_1 and R_4 so as to stay within the ratings of the device.

Over current faults can be cleared by combining the tunnel diode-transistor circuit with an SCR and a circuit breaker as shown in Figure 8-2-10.

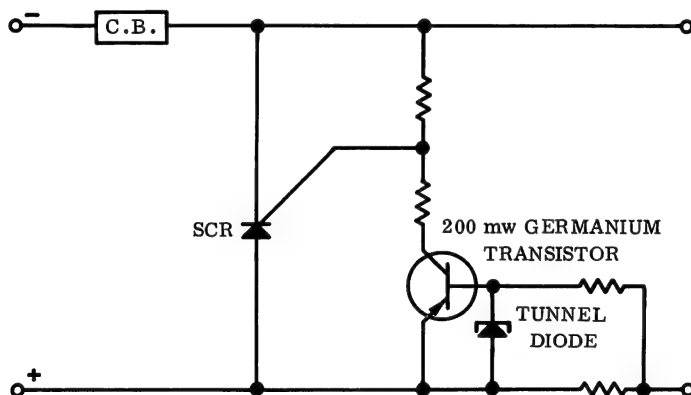


Figure 8-2-10 SCR Crow Bar Over-current Protection Circuit

SECTION 8-3

The Alternator Voltage Regulator

Another type of voltage regulator is the type used to control the charging rate of batteries, such as an alternator voltage regulator. Electromechanical devices have been employed for many years as voltage regulators, however, these devices tend to wear out. A solid state regulator will give greater reliability.

CIRCUIT DESCRIPTION

A schematic of a simplified solid state voltage regulator is shown in Figure 8-3-1.

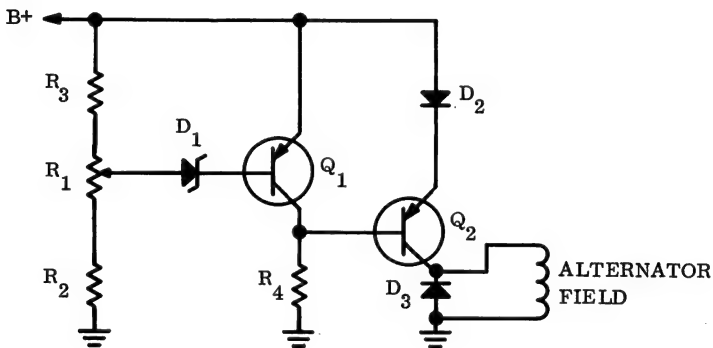


Figure 8-3-1 Simplified Solid State Voltage Regulator

The circuit operation is as follows:

The zener diode D_1 in conjunction with Q_1 is a voltage sensing circuit, turning the field switching transistor Q_2 , on when the battery is undercharged and off when the battery charge has attained the proper level. Under full charge conditions, Q_2 is cut off representing essentially an open circuit. There will be no current flow through the alternator field coil and the alternator will produce no charging current. As the battery voltage decreases, Q_2 is turned on so it represents a virtual short circuit between the field coil and the battery. During this time the alternator field is energized and a charging current is supplied to the battery. Diode D_3 is a field suppressor which prevents build-up of high induced voltage across the coil when the current through the coil is interrupted. Diode D_2 provides a reverse bias to reduce the leakage current of Q_2 during its off condition.

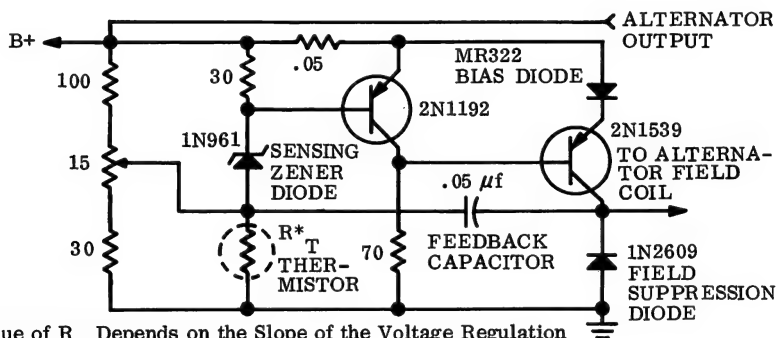
In the sensing circuit, D_1 conducts only when the battery voltage has reached the desired, full charge value. When the battery voltage is low,

there is no current flow through D_1 or the base-emitter junction of Q_1 which keeps Q_1 cut off. A sufficient voltage appears across this collector-to-emitter of Q_1 to produce a current flow in the base of Q_2 causing this transistor to turn on. This allows current flow through the field coil producing an alternator output to charge the battery.

When the desired charge is reached, D_1 conducts causing Q_1 to turn on which effectively shorts out the base-emitter junction of Q_2 . This cuts off Q_2 and turns off the current that was flowing in the field coil which reduces the output of the alternator.

In actual operation this switching action occurs many times each second, depending on the current drain from the battery. The battery charge, therefore, remains essentially constant and at the maximum value for optimum operation.

A schematic of a complete alternator voltage regulator is shown in Figure 8-3-2.



*The Value of R_T Depends on the Slope of the Voltage Regulation versus Temperature Curve.

Figure 8-3-2 Complete Solid State Alternator Voltage Regulator

SECTION 8-4

SCR Alternator Regulators

An alternator voltage regulator constructed from semiconductor devices has a number of advantages over the conventional electromechanical regulators. The SCR voltage regulator circuit shown in Figure 8-4-1 can accurately control the charge to the battery while offering the additional advantage of operating at higher temperatures than is possible in germanium transistor designs.

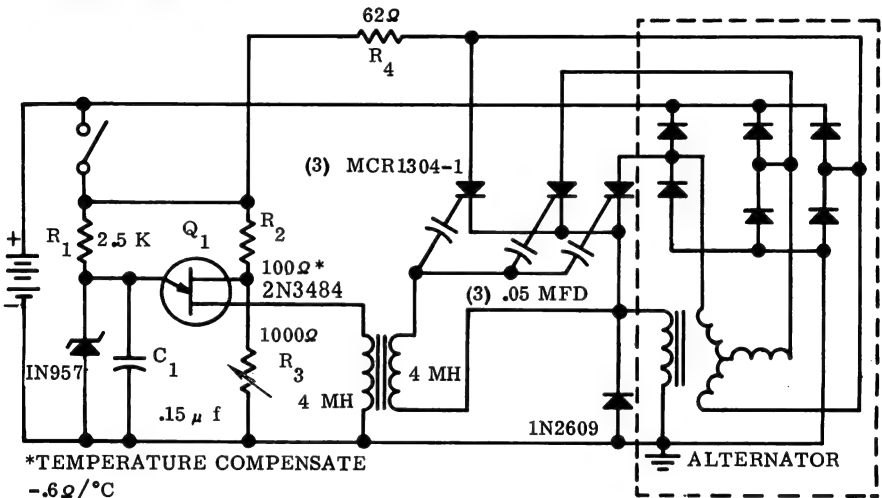


Figure 8-4-1 (3) SCR (3) Phase Alternator Voltage Regulator

In this circuit, the controlled rectifiers, when triggered, permit current flow through the alternator field winding during the positive excursions of the output from each of the three alternator field windings. Triggering is provided by a relaxation oscillator consisting of unijunction transistor Q_1 , a frequency determining network, R_1C_1 , and a voltage sensing circuit consisting of a zener diode and resistors R_2R_3 . The oscillator frequency is high compared to the alternator frequency to assure triggering during each positive excursion of the alternator waveform. When the waveform swings negative, the controlled rectifiers are turned off.

For Q_1 to oscillate, however, the voltage between B_1 and B_2 must be equal to or less than $\frac{V_{EB1}}{\eta}$: where η (intrinsic standoff ratio) is as specified on the unijunction data sheet. Since $\frac{V_{EB1}}{\eta}(\max)$ is determined by the zener diode clamping action and V_{BB} is determined by the voltage of the battery, oscillations and subsequent triggering will occur only when the battery voltage has dropped below the predetermined level set by R_2 and R_3 . Resistor R_4 provides initial start latching current for the SCR.

If the base-two voltage level is clamped by a low impedance source, it is only necessary to temperature compensate the changes in the zener and V_{EB1} junction to maintain constant voltage charging characteristics over a temperature range. If a charging voltage change with temperature is desired, additional temperature compensation will be necessary.

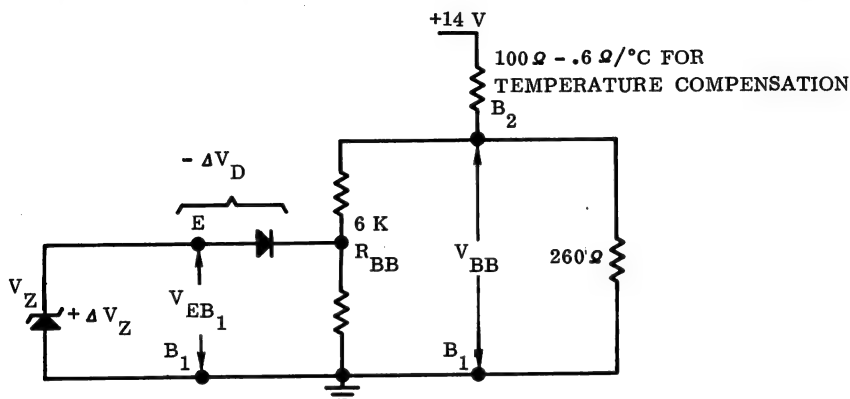


Figure 8-4-2 Basic Unijunction Voltage Sense Circuit

With reference to Figure 8-4-2, the required change in the unijunction base-to-base voltage is dependent upon the following relationship for proper temperature compensation.

$$\Delta V_{BB} = \frac{\Delta V_Z - \Delta V_D}{\eta}$$

Where

ΔV_Z = change in zener voltage level due to temperature.

ΔV_D = change in emitter to base-one diode voltage due to temperature (-2 mv/ C).

ΔV_{BB} = required change in unijunction base-to-base voltage with temperature.

η = intrinsic standoff ratio.

If a 7-volt zener diode is used, it will have a positive 3.5 mV rise/degree C. When this is combined with negative 2 mV/degree C of V_D and divided by η (about .7) the required change of V_{BB} is +8 mV/degree C. Thus, the 100 ohm resistor must have a T.C. of -.3 ohm/degree C. To cause a taper charge of -10 mV/degree C the T.C. of the 100 ohm resistor must be increased to -.6 ohm/degree C.

Figure 8-4-3 shows the use of a single SCR in series with two diodes to control the field current from only two phases of the alternator. It is apparent that the average current through the field will be reduced unless a lower field resistance is used. The output ripple of the alternator will also be greater.

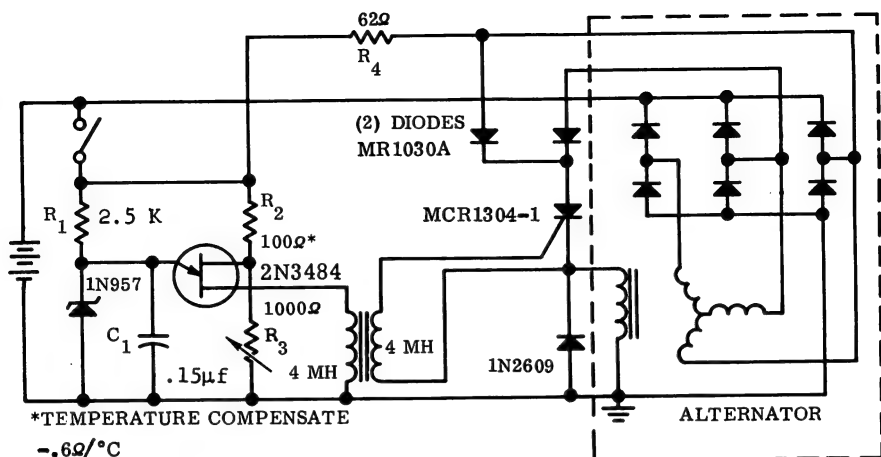


Figure 8-4-3 (1) SCR (2) Phase Alternator Voltage Regulator

It is also possible to operate from one phase of the alternator with a single SCR. This circuit would be economical as far as the regulator is concerned but copper and iron requirements in the alternator are increased to obtain the same output.

In the outboard engine field, the alternators may use permanent magnet fields. This increases the problem of regulating the output by limiting the control function to opening or shorting the output. Because of the high reactance source of most alternators, opening the output circuit will generally stress the bridge rectifiers to a very high voltage level. It is, therefore, apparent that the best control function would be shorting the output of the alternator for regulation of the charge to the battery.

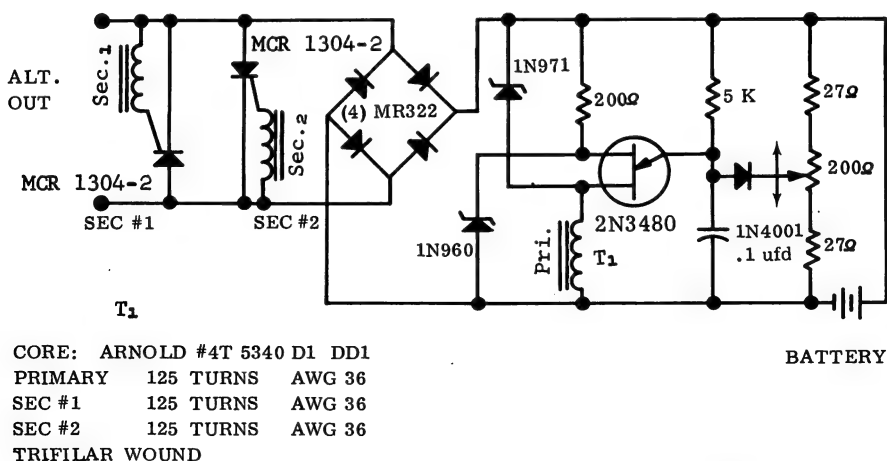


Figure 8-4-4 Permanent Magnet Field Alternator Regulator

Figure 8-4-4 shows a permanent magnet alternator regulator that was designed to regulate a 15 ampere output. The two SCR's are connected on the ac side of the bridge and short out the alternator when triggered by the unijunction voltage sensing circuit. In most cases the shorted output does not appreciably increase the maximum output current level.

A single SCR could be designed into the dc side of the bridge. However, because there is very little time for the SCR's to turn off at high alternator speeds, the turn-off requirements for the device are not in a practical range.

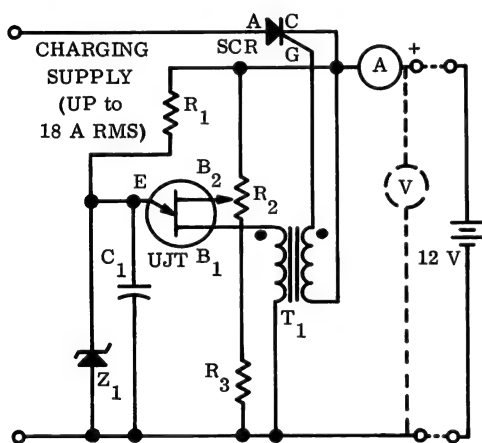
The unijunction circuit in Figure 8-4-4 will not oscillate until the input voltage level reaches the voltage determined by the intrinsic standoff ratio. The adjustable voltage divider will calibrate the circuit depending upon unijunction and zener tolerance. The series diode in the voltage divider circuit will compensate for the emitter base-one diode temperature change and temperature compensation of the voltage divider is necessary only for the zener diode temperature changes.

Because of the delay in charging the unijunction capacitor (Figure 8-4-4), if the battery is disconnected, the alternator voltage will rise to a high stress voltage on all components before the SCR's will be fired. The 1N971 zener was included in the circuit to provide a trigger pulse to the SCR's as soon as the alternator output voltage level approaches 30 volts.

SECTION 8-5

SCR Battery Charger

Reverse polarity battery connection, short circuiting the charger or overcharging the battery are often causes of battery or charger damage. A foolproof battery charger circuit is shown in Figure 8-5-1. This circuit will not operate until the battery, to be charged, is properly connected to the charger. The battery voltage controls the charger and will dictate its operation. When the battery is properly charged, the charger will cease operation.



Parts List

- R_1 3.9 K 1/2 W
- R_2 1 K Potentiometer
- R_3 5.1 K, 1/2 W
- C_1 0.25 μ f
- Z_1 1N753-6.2 V 400 mw Zener Diode
- SCR Motorola MCR808-3
- UJT 2N3480
- T_1 Primary 30T, #22
Secondary 45T #22
Core: Ferroxcube
203 F 181 - 3C3

Figure 8-5-1 12-Volt Battery Charger Control

The battery charging current is obtained through the controlled rectifier when it is triggered into the conducting state. Triggering pulses for the controlled rectifier are generated by a unijunction transistor relaxation oscillator (Figure 8-5-2). This oscillator is activated when the battery voltage is low.

While operating, the oscillator will produce pulses in the pulse transformer connected across the resistance, R_{GC} (R_{GC} represents the gate-to-cathode resistance of the controlled rectifier), at a frequency determined by the resistance, capacitance, R.C. time delay circuit.

Since the base-to-base voltage of the unijunction transistor is derived from the charging battery, it will increase as the battery charges. The increase in base-to-base voltage of the unijunction transistor causes its peak point voltage (switching voltage) to increase. These waveforms are sketched in Figure 8-5-2. (This voltage increase will tend to change the pulse repetition rate, but this is not important.)

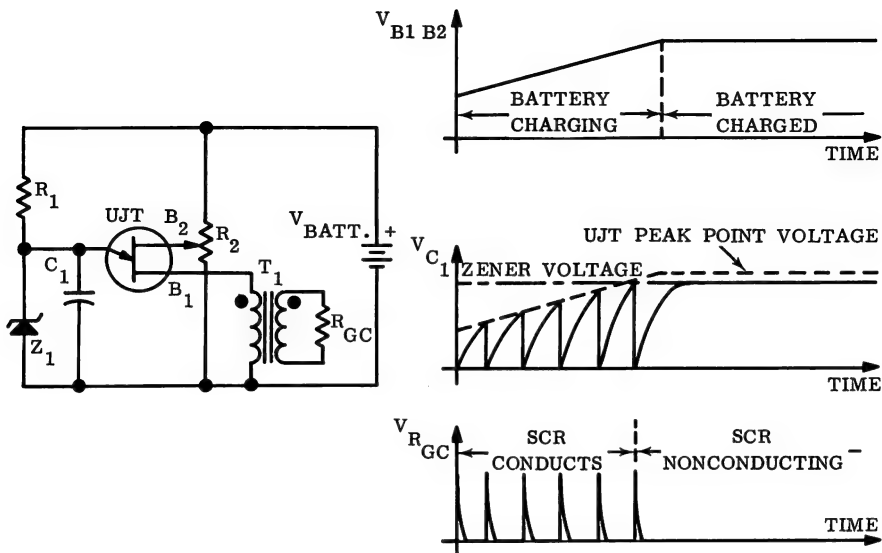


Figure 8-5-2 UJT Relaxation Oscillator Operation

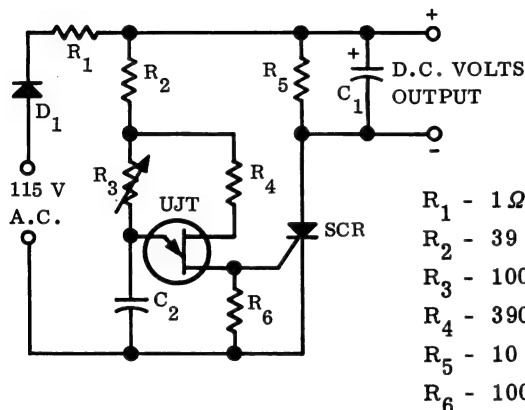
When the peak point voltage (switching voltage) of the unijunction transistor exceeds the breakdown voltage of the zener diode, Z₁, connected across the delay circuit capacitor, C₁, the unijunction transistor ceases to oscillate. If the relaxation oscillator does not operate, the controlled rectifier will not receive trigger pulses and will not conduct. This indicates that the battery has attained its desired charge as set by R₂.

The unijunction cannot oscillate unless a voltage somewhere between 3 volts and the cutoff setting is present at the output terminals with polarity as indicated. Therefore, the SCR cannot conduct under conditions of a short circuit, an open circuit, or a reverse polarity connection to the battery.

SECTION 8-6

SCR Transformerless DC Power Supply

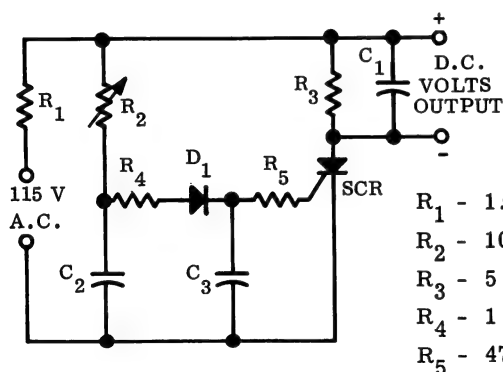
Lightweight dc power supplies are often required to operate transistorized radios or electronic circuits. The circuits, Figures 8-6-1 and 8-6-2, are two versions of a transformerless dc power supply. Both of these power supplies use phase shift circuits to control the firing angle of the controlled rectifier. The SCR firing angle determines the dc output voltage.



Parts List

R_1 - $1\ \Omega$ (1 W)	C_1 - $1000\ \mu f$ (15V)
R_2 - $39\ K\Omega$	C_2 - $0.5\ \mu f$
R_3 - $100\ K\Omega$	D_1 - 1N4003
R_4 - $390\ \Omega$	UJT - 2N3480
R_5 - $10\ K\Omega$	SCR - MCR1304-4
R_6 - $100\ \Omega$	

Figure 8-6-1 Transformerless D.C. Power Supply



Parts List

R_1 - $1\ \Omega$ (1W)	C_1 - $1000\ \mu f$ (15V)
R_2 - $10\ K\Omega$	C_2 - $1.5\ \mu f$
R_3 - $5\ K\Omega$	C_3 - $15\ \mu f$
R_4 - $1\ K\Omega$	D_1 - 1N4001
R_5 - $470\ \Omega$	SCR - MCR1304-4

Figure 8-6-2 Transformerless D.C. Power Supply

The phase delayed triggering of the power supply in Figure 8-6-1 is accomplished by unijunction transistor triggering. Resistors, R_2 and R_3 and capacitor C_2 determine the time delay of the unijunction switching. Increasing the value of resistor R_3 will decrease the output voltage.

Filtering of the chopped ac input is accomplished by capacitor C_1 . With the 1000 microfarad capacitor a 2.5 per cent ripple voltage appears across the load resistance when the supply is operated at its design value; 100 milliamperes, 10 volts. The total power supply efficiency, at the design value, is about 77 per cent.

The circuit in Figure 8-6-2 accomplishes phase delay from the R_5C_2 , R_4C_3 time delay circuit. Note that the unijunction transistor has been eliminated reducing the cost of the circuit.

Figure 8-6-3 compares the load regulation characteristics of the two circuits. The unijunction transistor triggered circuit is the better of the two at higher load currents. Both circuits were designed to deliver 100 milliamperes at 10 volts. Circuit optimization can improve the characteristics for higher currents.

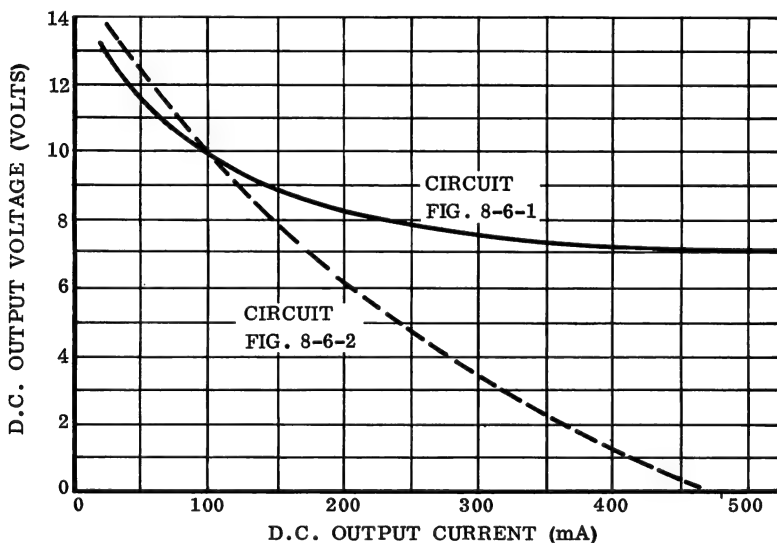


Figure 8-6-3 Load Regulation Characteristics
D.C. Transformerless Power Supply

SECTION 9-1

Automatic Vending Machine Circuit

A vending machine is required to perform a definite function after a series of operations have been performed by the operator. Above all, the circuit should never operate when no coin is inserted, and the machine must never give more than one piece of merchandise for each coin inserted. The circuit of Figure 9-1-1, will perform this prescribed function with the utmost accuracy and reliability.

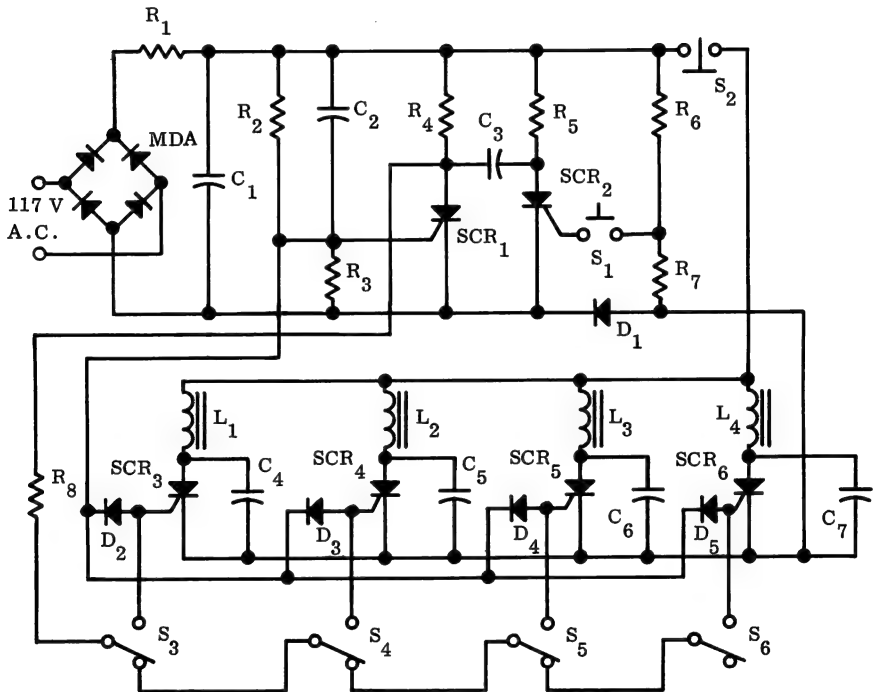


Figure 9-1-1 Automatic Vending Circuit

A description of the circuit's operation can best be accomplished by identifying and discussing the circuit functions as shown in the block diagram of Figure 9-1-2. As the discussion proceeds, note that many circuit components are used to perform more than one circuit function.

The circuitry in the "Coin Operated Activating Circuit" activates the vending machine when the proper coin is inserted. But before we insert the coin, let's examine the circuit under static conditions.

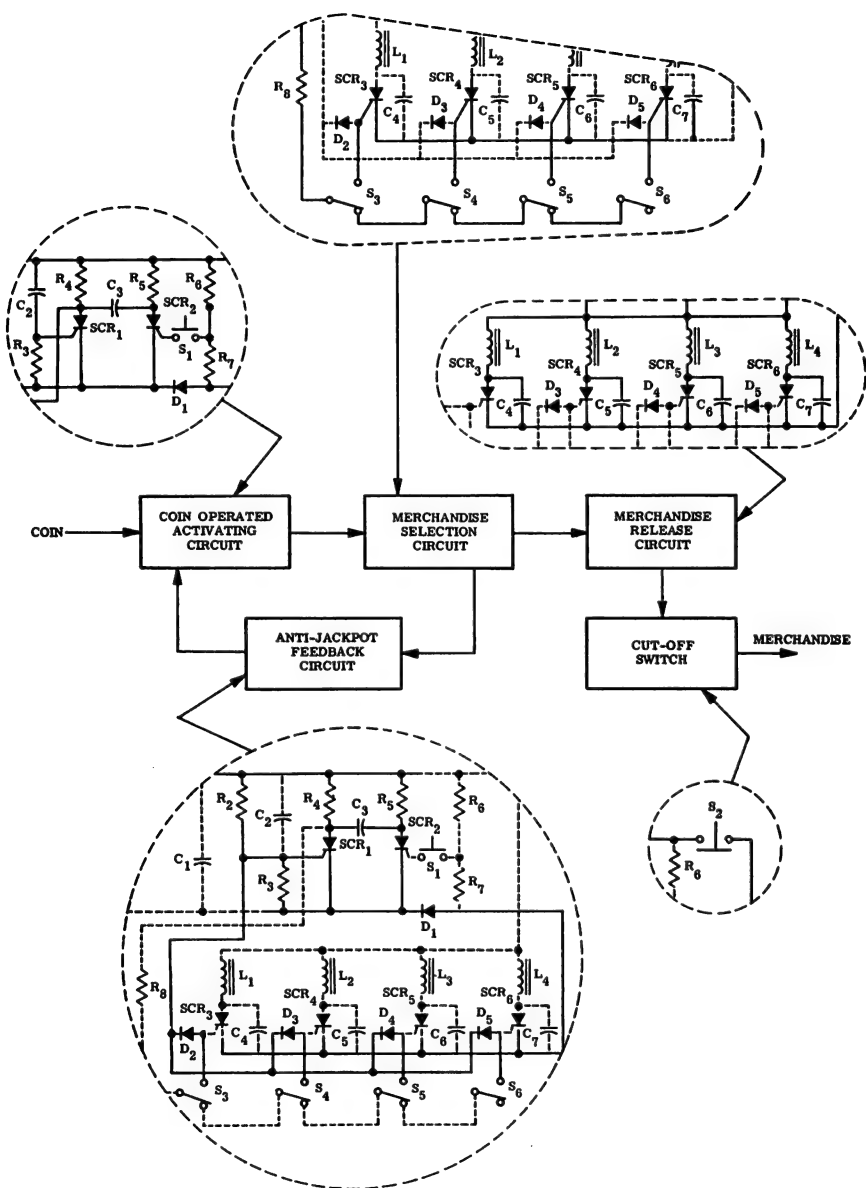


Figure 9-1-2 Block Diagram

A dc potential obtained from the rectified ac line voltage, in the sketch, appears across the uppermost and the lowermost horizontal lines; the upper line being positive with respect to the lower. Capacitor C_2 has charged to this dc potential through resistor R_3 . At the instant of charging, the gate-to-cathode terminals of SCR_1 reach triggering conditions and cause SCR_1 to go into conduction. SCR_2 remains non-conducting. The capacitor C_2 is charged to approximately the dc potential with its right terminal positive with respect to the left terminal. Therefore, in the static condition SCR_1 is conducting, SCR_2 is non-conducting and C_3 is charged.

When the proper coin is inserted, switch S1 closes, applying a trigger voltage to SCR_2 , causing it to conduct. The discharge of capacitor C_2 through SCR_2 and SCR_1 , causes SCR_1 to be back biased and become non-conducting. Now, SCR_1 is non-conducting and SCR_2 is conducting. Capacitor C_2 is now charged with the left terminal positive with respect to the right terminal.

The off condition of SCR_1 places a positive potential on the connection going to the selection switch (S_3 , S_4 , S_5 and S_6). Manual activation of a selection switch connects a triggering potential to the proper controlled rectifier circuit, SCR_1 , SCR_2 , SCR_3 , or SCR_4 .

The "Merchandise Release Circuit" includes the triggered controlled rectifier in series with a vending solenoid which mechanically releases the selected merchandise when activated. Capacitors C_4 through C_7 are included to prevent dv/dt triggering of the controlled rectifiers if any fast rising pulses are caused by the circuit operation.

The operation of a selector switch, in the process of triggering the selector controlled rectifier, also connects the trigger signal, through the SCR's gate diode, D_2 , D_3 , D_4 or D_5 , to the gate of SCR_1 . Thus, this mechanical selection operation also causes SCR_1 to conduct, reducing the voltage applied from the anode of SCR_1 to the selector switches to about one volt. Since a one volt gate-to-cathode potential may trigger the selector controlled rectifiers, the effective gate-to-cathode potential is reduced to about zero by placing the diode D_1 in the current return path of these SCR's. Thus, in the time it takes an SCR to go into conduction (about 1 to 2 microseconds) the selection potential is removed and the circuit reverts to its disarmed status, ready for the next coin operation.

The turn-off of the operating solenoid is achieved by turning off the conducting SCR. This mechanical operation is accomplished by opening the operating current path with a merchandise activated "cut-off switch."

It is feasible that this type of circuit operation may be used in certain manufacturing control processes.

SECTION 9-2

Process Control

The development of resistive elements whose resistance value is sensitive to changes in solution "ph", humidity, temperature, pressure, light intensity, etc. has lead to the design of circuitry to automatically control these parameters. Figure 9-2-1 shows the circuit diagram of such a control circuit.

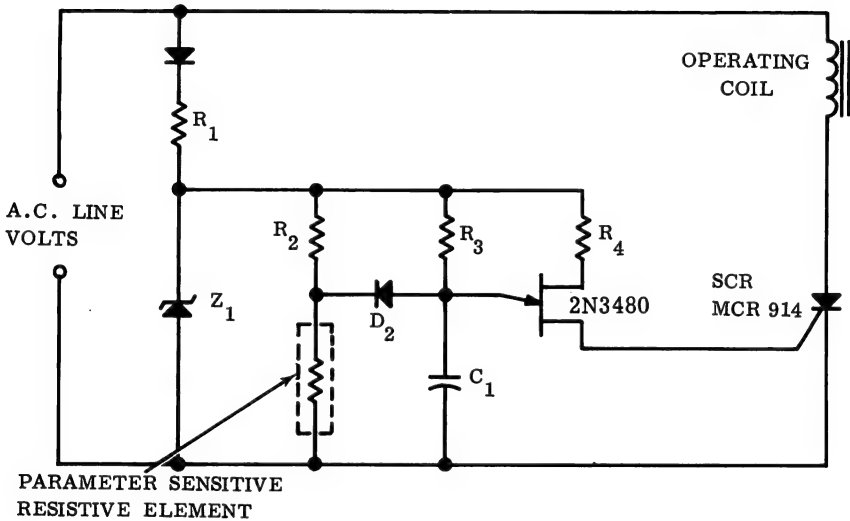


Figure 9-2-1 Process Control Circuit

Originally, the circuit was developed to control the soap concentration of a dishwashing solution. The controlling resistor is sensitive to the solution "ph" factor. A decrease in soap concentration causes the circuit to operate a solenoid valve, adding soap until the desired concentration is attained.

The controlled rectifier triggering is obtained from a unijunction oscillator (Figure 9-2-2), operating by charging and discharging capacitor, C_1 through resistor, R_3 , and the unijunction, respectively. Note that the trigger pulses are transferred to the controlled rectifier gate-to-cathode terminals (represented by R_{GC} in Figure 9-2-2) only when the unijunction peak point voltage is developed across the capacitor C_1 .

In the actual control circuit (Figure 9-2-1) the resistor capacitor charging circuit R_3C_1 , is bridged by a diode, D_2 , with its cathode connected to a voltage divider. (This is detailed in Figure 9-2-3.) The capacitor, C_1 , can now only charge to about one volt higher than the potential set by the divider; one section of which is the parameter sensitive resistance element.

As the resistance of this element increases due to solution “ph” change, the voltage across the capacitor, C_1 , increases. When the voltage across the capacitor reaches the unijunction peak point voltage, the oscillator will operate and the controlled rectifier will trigger.

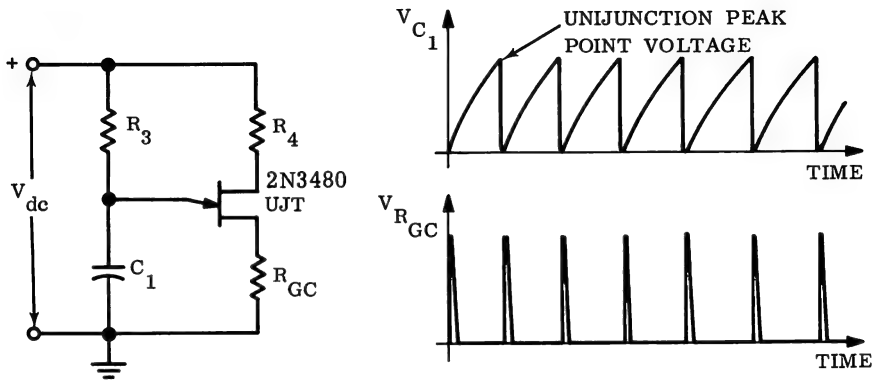


Figure 9-2-2 Basic Unijunction Oscillator Circuit

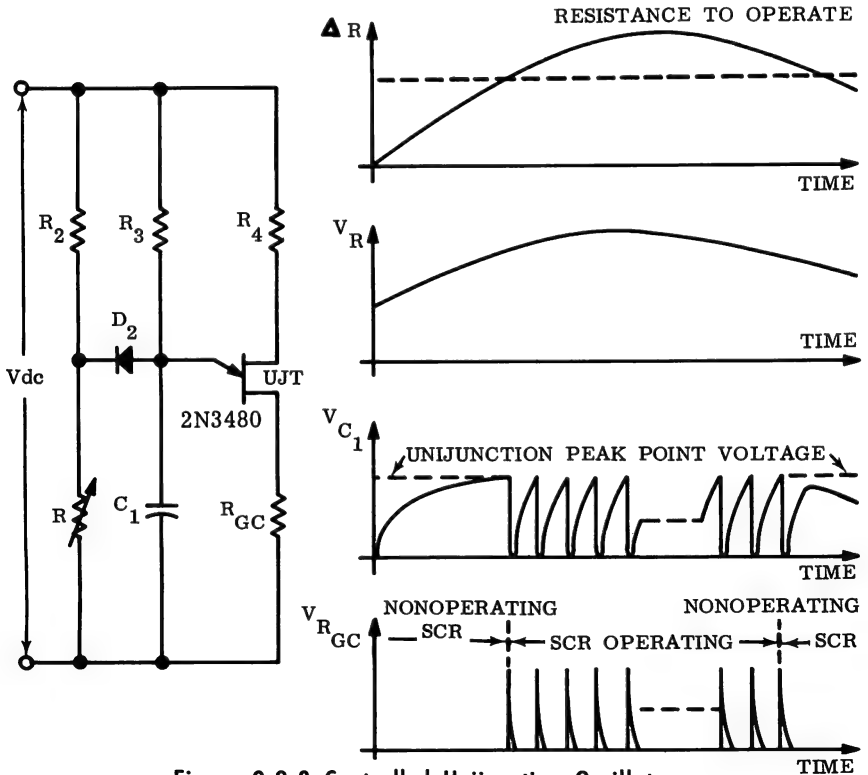


Figure 9-2-3 Controlled Unijunction Oscillator

The triggering of the controlled rectifier also operates the operating coil, (solenoid valve in the soap solution control), since both are connected in a series circuit.

Now the resistor starts to sense the solution concentration increase by decreasing in resistance value. When it reaches the set resistance value, the circuit reverts to its non-operating state. The controlled rectifier turns off and the valve closes.

This control circuit has been discussed as a solution control but as mentioned above, it can control humidity, temperature, pressure, light intensity and any other controllable parameter by using the proper resistive sensing element and the proper operating coil.

SECTION 9-3

SCR Ring Counter

Igmor, the stone age man, found it extremely difficult to count above twenty. That is if he were normal and had only ten fingers and ten toes. The present day Igmor is not only interested in counting numbers raised to millions of powers but he wants to know how fast and to what accuracy he can do the counting.

Counting (and indicating the counts if desired) is the only purpose of the circuit sketched in Figure 9-3-1. It can be used to count the number of milliseconds in a 24 hour day or the number of peas in a can.

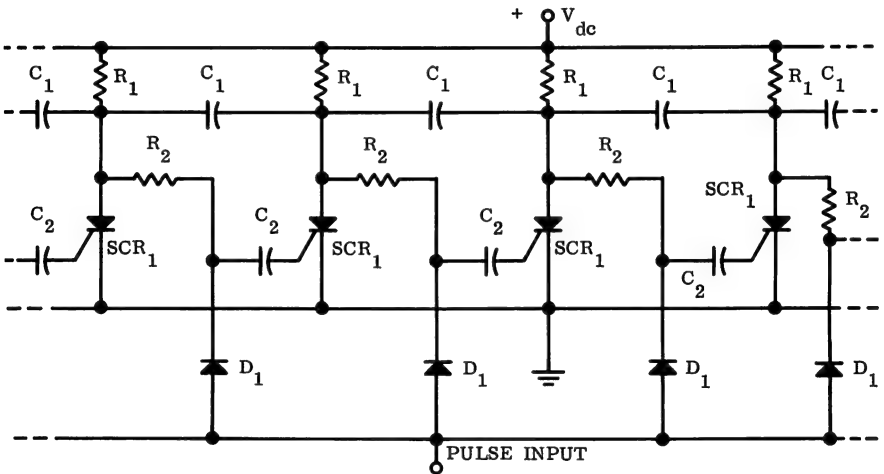
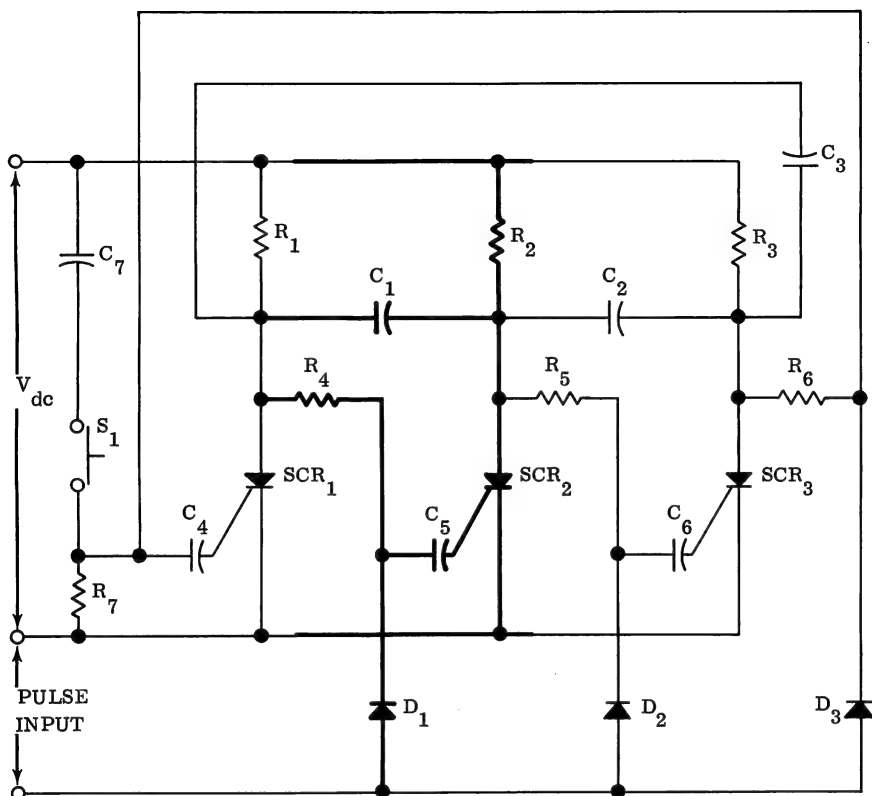


Figure 9-3-1 SCR Counter String

The counting function is achieved by propagating the conduction, (on), state along a string of SCR counting stages. Thus, the SCR which is conducting represents the count above the first stage in the string. The propagation vehicle is a short duration voltage pulse applied to the pulse input terminals. These pulses may be derived from a variety of pulse producing circuits that are unique to the desired application.

The counter's operation and characteristics can be illustrated using the Three Stage Ring Counter shown in Figure 9-3-2.

The portion of the circuit emphasized by bold lines is one basic stage of the counter. Note that each stage requires only two semiconductor components and four other electrical components.



TYPICAL VALUES:

$R_1, R_2, R_3 - 1\text{ K}\Omega$

$R_4, R_5, R_6 - 22\text{ K}\Omega$

$R_7 - 5\text{ K}\Omega$

$C_1, C_2, C_3 - 0.2\text{ }\mu\text{fd}$

$C_4, C_5, C_6 - 0.001\text{ }\mu\text{fd}$

$D_1, D_2, D_3 - 1\text{N}4002$

$\text{SCR}_1, \text{SCR}_2, \text{SCR}_3 - \text{MCR}914\text{-}3$

$S_1 - \text{SPST PUSH BUTTON}$

$V_{\text{dc}} - 100\text{ V}$

Figure 9-3-2 Three Stage Ring Counter

A step-by-step discussion of the circuit operation is useful in pointing out some basic SCR operating characteristics as well as determining circuit operation. Let's assume that the circuit components have the values listed as typical in the sketch.

Initially the circuit is static with V_{dc} (100 volts) appearing from anode-to-cathode of all the controlled rectifiers. The closing of the switch, S_1 , applies a positive voltage pulse from gate-to-cathode of SCR_1 , causing it to conduct a current through resistor R_1 . SCR_1 now has an anode-to-cathode potential equal to its conducting voltage drop (about 1 volt).

Since SCR_2 and SCR_3 are not conducting, their anode-to-cathode potentials remain at 100 volts. Thus, capacitor C_1 becomes charged to about 99 volts; the terminal connected to SCR_2 being positive with respect to the terminal connected to SCR_1 . The low conducting potential across SCR_1 also allows diode D_1 , through resistor R_4 , to be forward biased to any pulse input greater than two volts. Note that diodes D_2 and D_3 require potentials greater than about 101 volts to become forward biased.

The application of a single input pulse greater than 2 volts but less than 101 volts (normally from 5 to 10 volts) will be transmitted through capacitor C_5 , to the gate-to-cathode terminal pair of SCR_2 . This trigger pulse causes SCR_2 to quickly go into conduction, effectively connecting the charged capacitor C_1 directly across SCR_1 . As capacitor C_1 discharges through SCR_2 (in its conducting direction) and SCR_1 (in its reverse direction), the controlled rectifier SCR_1 reverts to its non-conducting or blocking state. Thus, we have propagated the conducting state from SCR_1 to SCR_2 . The above described analysis now applies to the components connected to SCR_2 rather than SCR_1 . Each successive input pulse will turn off the conducting controlled rectifier and turn on the following controlled rectifier.

Figure 9-3-3 illustrates the waveforms that appear on the designated circuit terminals.

The described counting function may be used as the basic or auxiliary building blocks in a number of counting circuits. Figure 9-3-4 illustrates in block form a zero-to-99 counter. Note that only twenty counting stages are required to perform this count, the first ten stage counter indicating the units digits and the second ten stage counter indicating the tens digits. A zero-to-999 counter requires only 30 stages, a zero-to-9999 counter requires only 40 stages, etc.

The interstage coupling circuit shown in Figure 9-3-5 requires only a few components. It is usually connected to the final counting stage (numerically nine, in a ten stage counter). Thus, at the turn on of the initial stage (numerically zero), a pulse is transmitted to the next stage counter.

An often useful application of the counter is a pulse frequency divider shown in Figure 9-3-6. Here an "n" stage counter is used to produce an output pulse frequency equal to the input frequency divided by the number of counting stages; ie:

$$f_{out} = f_{in}/n$$

Slow accurate repetition rates are easily produced using this approach.

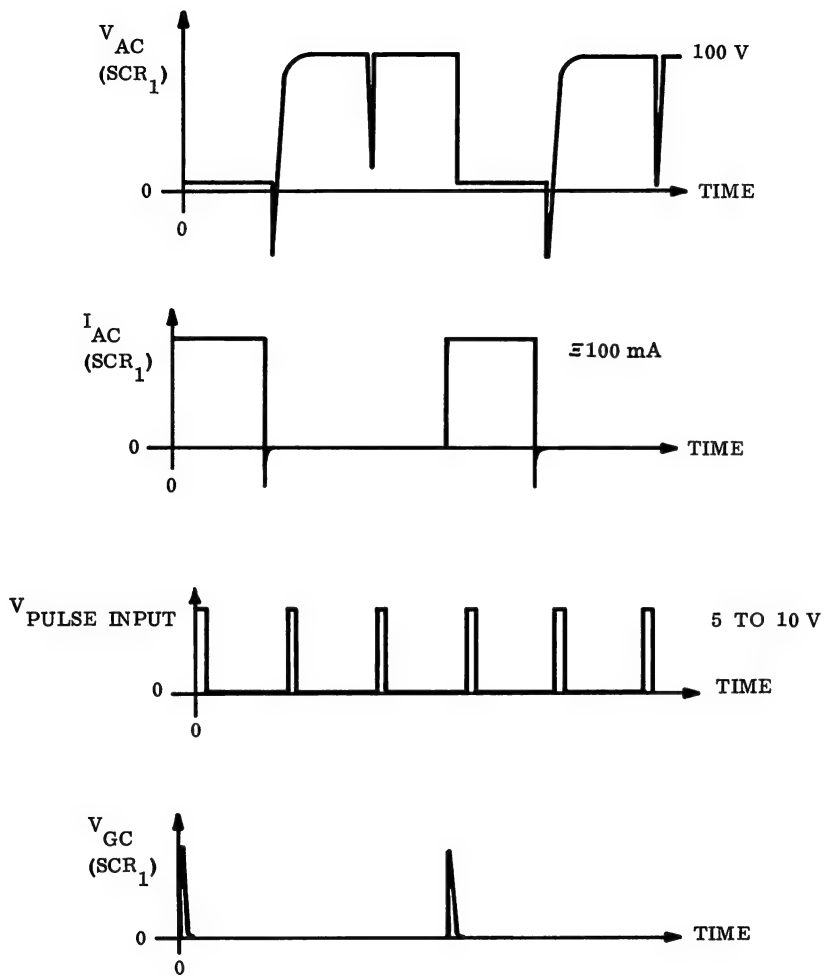


Figure 9-3-3 Three Stage Ring Counter Waveforms

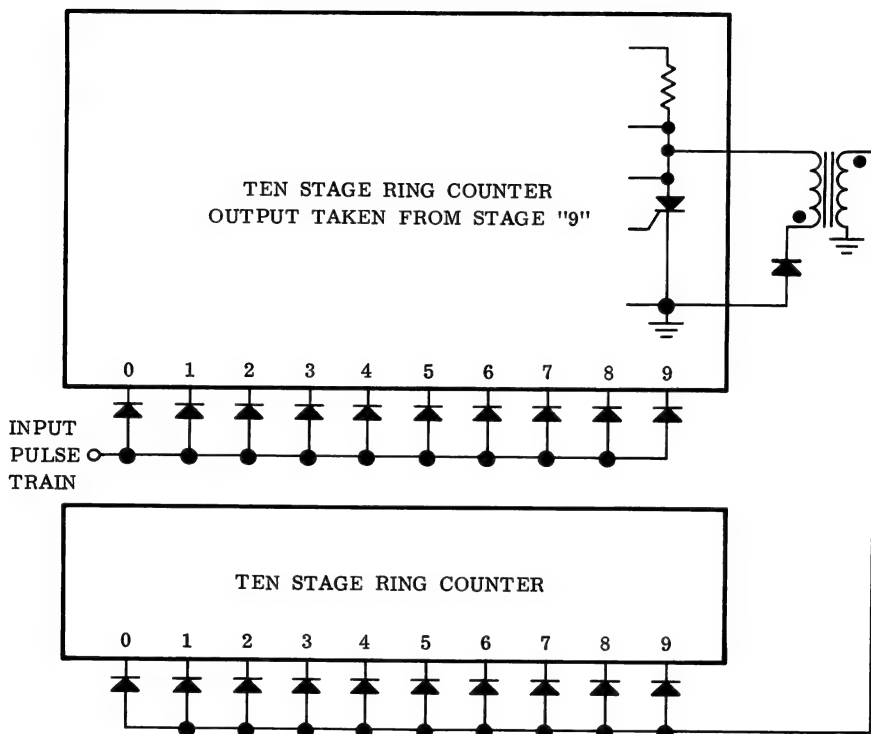


Figure 9-3-4 Zero-to-99 Counter

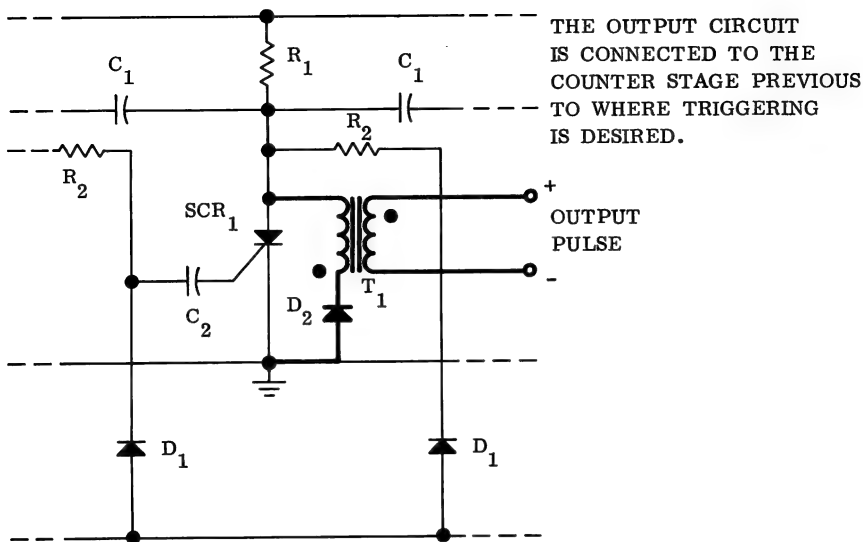
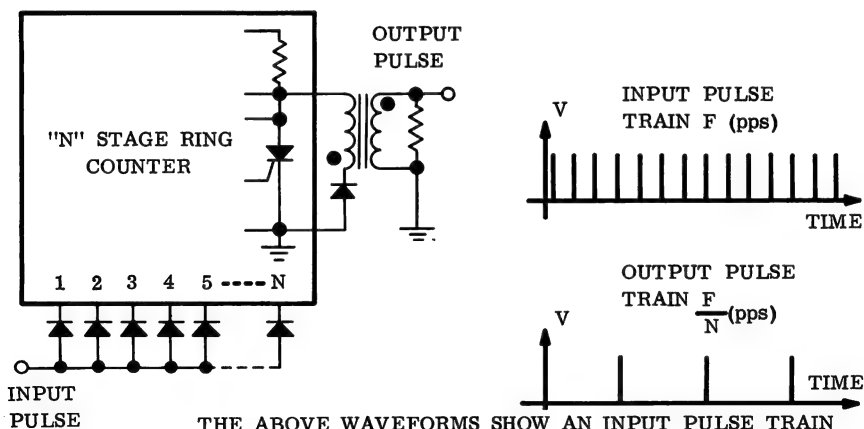


Figure 9-3-5 Output Circuit for Counter



THE ABOVE WAVEFORMS SHOW AN INPUT PULSE TRAIN OF FREQUENCY, F (pps). USING A 4 STAGE COUNTER AS ILLUSTRATED IN THE WAVEFORM WE OBTAIN AN OUTPUT PULSE TRAIN OF FREQUENCY, $F/4$ (pps).

Figure 9-3-6 Frequency Divider

SCR counter circuits are useful in marine depth finders, ballistic missile detonator timers, and LORAN aircraft navigation indicators. Block diagrams, Figures 9-3-7 through 9-3-9 respectively, illustrate these applications.

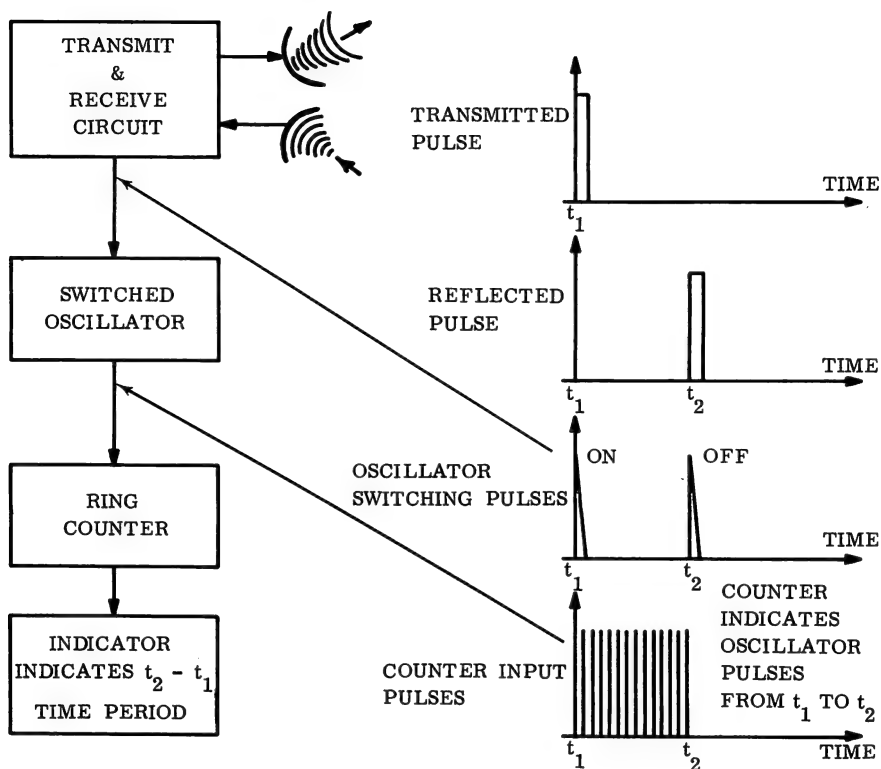


Figure 9-3-7 Marine Depth Finder

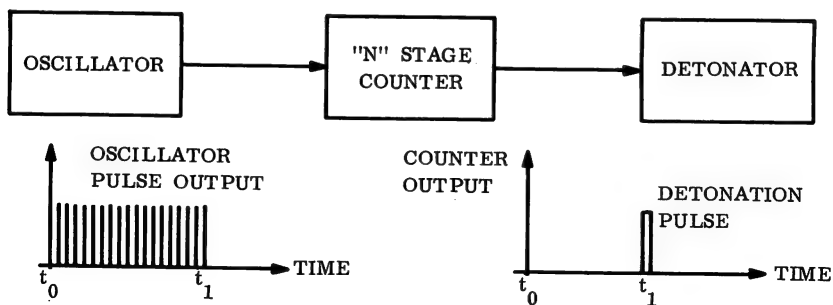


Figure 9-3-8 Ballistic Missile Detonator Timer

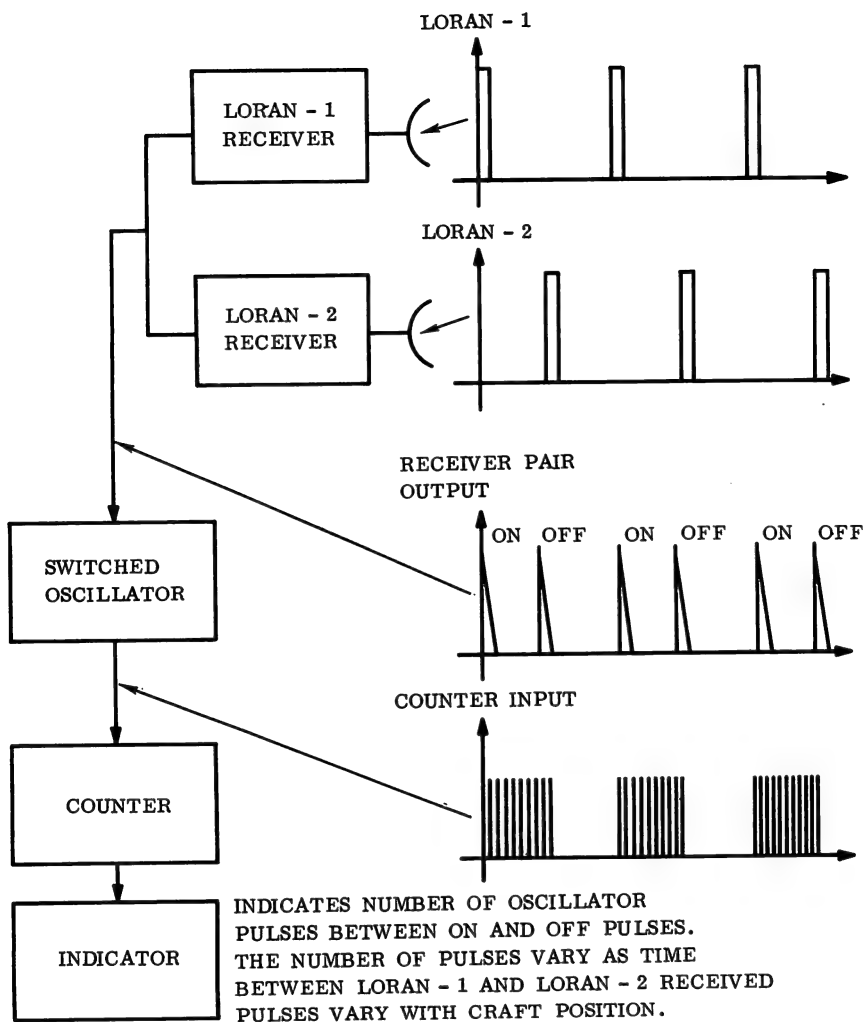
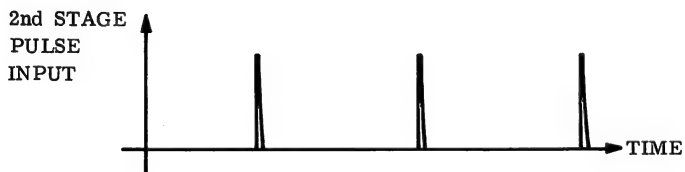


Figure 9-3-9 Loran Aircraft Navigation Timer

A counter whose operation is similar to that of the ring counter discussed previously is a simple Binary Counter circuit. Two stages of the counter are illustrated in Figure 9-3-10.



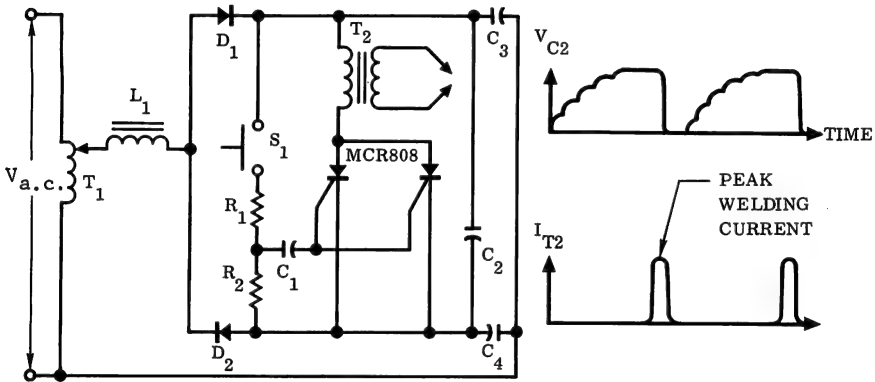
9-3-9

SECTION 9-4

Controlled Rectifier Welding Controls

The control of high currents for short periods of time, as required in spot welding, can easily be accomplished by high current controlled rectifiers. Described below are two basic types of controlled rectifier welder circuits; 1) the capacitor discharge circuit, and 2) the ac burst control circuit.

Figure 9-4-1 illustrates the capacitor discharge welder circuit operation from the ac line voltage. A high dc potential is developed across the capacitive voltage doubler, consisting of capacitors C_2 , C_3 and C_4 . The charge stored will depend on the total capacitance value and the available voltage, as set by the auto-transformer.



Parts List

R_1 - 2 K	D_1 - 25 A (200V) MR324
R_2 - 1 K	D_2 - 25 A (200V) MR324
C_1 - .1 μ f(150V)	T_1 - Auto transformer
C_2 - 2000 μ f(400V)	T_2 - Welding transformer
C_3 - 2000 μ f(400V)	L_1 - Choke
C_4 - 2000 μ f(400V)	S_1 - Push switch

Figure 9-4-1 Capacitor Discharge Welder

Welding currents are conducted in the secondary of the transformer, T_2 , as the capacitors are discharged through the controlled rectifiers, SCR_1 and SCR_2 , and the transformer primary. This occurs when the triggering switch, S_1 is operated.

The circuit waveforms as seen on an oscilloscope connected across the indicated terminals are also sketched in Figure 9-4-1. Note that the MCR808 controlled rectifiers used in this circuit must pass extremely high surge currents. One MCR808 would be adequate in this circuit, but two or more may be used to decrease the discharge path resistance, yielding higher output currents.

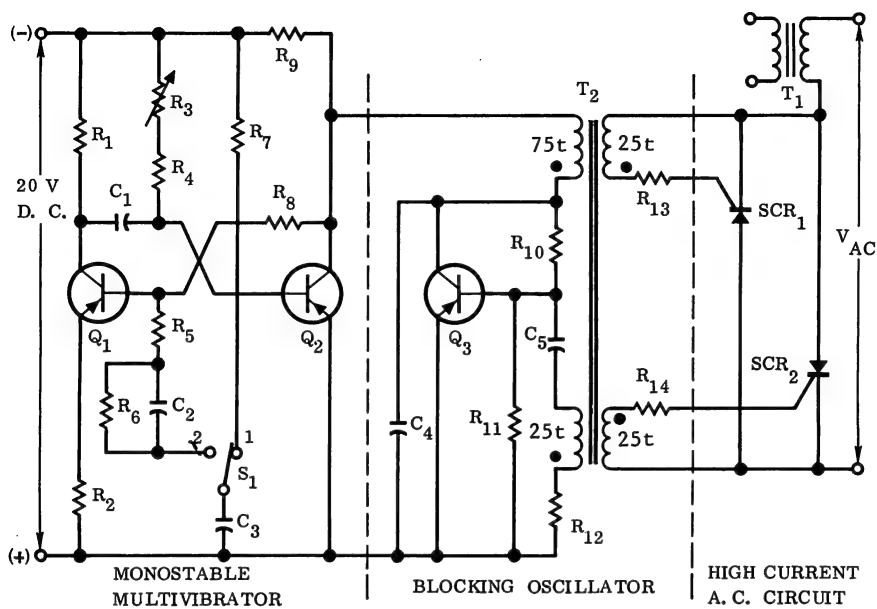
The other welder circuit, Figure 9-4-2 controls the time duration of an ac line voltage applied to the welding transformer, T_1 .

Controlled rectifier trigger pulses of about two microseconds duration are produced by the blocking oscillator, operating at about 10,000 pulses per second. The blocking oscillator is controlled by the monostable multivibrator.

Initially transistor Q_1 is cut off, transistor Q_2 is saturated, and switch S_1 is in position one; allowing capacitor C_3 to charge to 20 volts. The saturated condition of transistor Q_2 keeps the blocking oscillator circuit connected across a low potential and it does not oscillate.

When switch, S_1 , is operated, (switched to position 2), the charged capacitor, C_3 , causes transistor, Q_1 , to go into conduction and cut-off transistor Q_2 , through capacitor, C_1 . Now the blocking oscillator can oscillate until capacitor, C_1 , charges through resistors R_3 and R_4 and turns on transistor, Q_2 . Thus, the number of ac cycles in a welding burst is set by capacitor, C_1 and resistors R_3 and R_4 .

The practical time control range is from 20 milliseconds to one second.



R_1 - 1.2 K Ω
 R_2 - 300 Ω
 R_3 - 1 K Ω
 R_4 - 3 K Ω
 R_5 - 240 Ω
 R_6 - 3 K Ω
 R_7 - 5.1 K Ω
 R_8 - 5.1 K Ω

R_9 - 180 Ω
 R_{10} - 68 K Ω
 R_{11} - 10 K Ω
 R_{12} - 100 Ω
 R_{13} - 10 Ω
 R_{14} - 10 Ω
 C_1 - 50 μ fd
 C_2 - 0.1 μ fd

C_3 - 0.5 μ fd
 C_4 - 240 μ fd
 C_5 - 0.01 μ fd
 Q_1 - 2N650
 Q_2 - 2N650
 Q_3 - 2N651
 SCR_1 - 2N2577
 SCR_2 - 2N2577

T_1 - Weld trans.
 T_2 - Arnold Core #4T5651D1 turns as indicated on sketch above.

Figure 9-4-2 A-C Burst Welder

SECTION 10 Lamp Control

SECTION 10-1 Vehicular Flasher

The flasher described in this circuit was particularly designed for aircraft operation, where severe voltage transient conditions make transistor circuits prohibitively expensive. However, this type of flasher circuit is also attractive for marine use, since it is not as vulnerable to the corrosive affects of high humidity, or salt atmosphere as are metallic contacts. The elimination of metallic contacts make it attractive for use wherever a reliable flasher is required.

Basically, the circuit is similar to many other SCR flasher circuits which make use of the familiar flip-flop. However, there is one important difference to insure stable and reliable operation. In the basic flip-flop, SCR₁ and SCR₂ alternately latch on, and in conjunction with the commutating capacitor, C₁, turn the other off.

With only the basic circuit, a voltage transient, a momentary load short circuit, or some other disturbance could cause both SCR's to turn on at the same time. When this occurs, the circuit ceases to operate and it cannot resume operation until the supply voltage is momentarily interrupted. In the circuit of Figure 1, resistor R₂ is made large enough so that the current which it supplies to SCR₁ is always less than the holding current. Therefore, SCR₁ can never remain in conduction. This improvement also eliminates the need for elaborate starting circuitry.

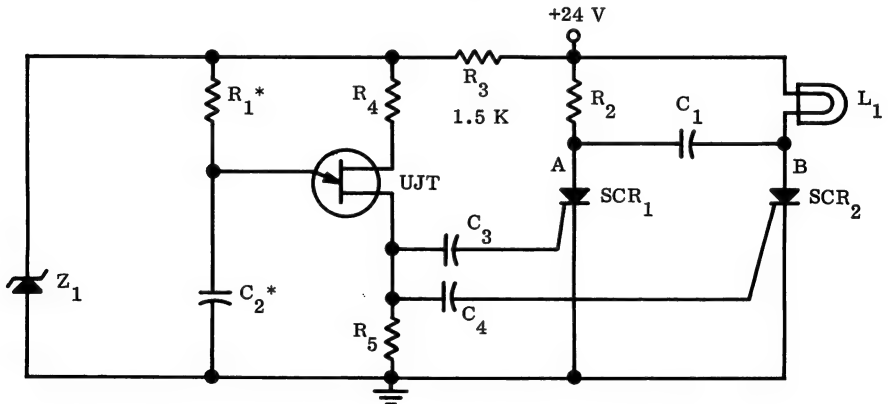


Figure 10-1-1 Vehicular Flasher

The circuit operates as follows: The unijunction transistor operates as a stable relaxation oscillator which produces trigger pulses for both

SCR's at a rate equal to twice the desired flash rate. At the start of operation, both SCR's attempt to turn on. SCR₁ is not able to latch on because its anode current is limited by R₂ and it never reaches the holding current level. With SCR₂ conducting, Point B is reduced to a voltage level which is above ground by an amount equal to the forward drop of SCR₂. C₁ charges through R₂ with Point A rising to the supply voltage. When the next trigger pulse occurs, SCR₁ can turn on since it is supplied with anode current by C₁. With SCR₁ in conduction, SCR₂ is effectively reverse biased by C₁, which causes it to turn off. The load current is supported by C₁ for a few microseconds during the turn off interval of SCR₂. SCR₁ continues to remain in conduction as C₁ charges through the lamp with Point B rising to the supply voltage. When the sum of the charging current of C₁ and the current through R₂ drops to a value below the holding current of SCR₁, SCR₁ drops out of conduction. C₁ then discharges through R₂ and the lamp at a low current level. When the next trigger pulse arrives, the cycle is repeated.

Note that when SCR₁ discharges C₁ to turn off SCR₂, SCR₂ is also being supplied with a turn-on trigger pulse. The trigger pulse must be kept short in duration in comparison with the turn off pulse supplied by C₁, in order to insure that SCR₂ will turn off. If for some reason, SCR₂ should fail to turn off as intended, C₁ will recharge with Point A positive with respect to Point B, and thus turn off pulses will be supplied until it does turn off. Likewise, if one of the SCR's should fail to turn on, turn-on pulses will continue to be applied until it does turn on. Since the firing voltage of the unijunction transistor is not voltage sensitive, the repetition rate of the trigger pulses are not affected by voltage changes. The unijunction trigger also makes the flash rate insensitive to temperature changes.

TABLE 10-1-1

Parts List for Figure 10-1-1

R ₁	910K, 1/2 W	C ₂	0.5 μ fd	UJT 2N3480
R ₂	12K	C ₃	0.1 μ fd	Z ₁ 1N971A
R ₃	1.5K, 1/2 W	C ₄	0.1 μ fd	27V, 400 mw
R ₄	470 Ω	SCR ₁	Motorola	L ₁ 312
R ₅	47 Ω		MCR 808-3	
C ₁	2 μ fd	SCR ₂	Motorola	
			MCR 808-3	

* R₁, C₂, and the unijunction transistor intrinsic standoff ratio determine repetition rate. Values shown give a rate of approximately 70 flashes per minute.

SECTION 10-2

Incandescent Lamp Dimmers

The SCR phase control circuit has made feasible a high capacity lamp dimmer which is compact enough to fit in the space normally provided for mounting of a toggle switch. In order to be attractive for domestic use, such a dimmer must be low in cost, and yet must provide reasonable stability and reliability. The use of a neon bulb for triggering permits the design of circuits which are less susceptible to temperature drift than the phase shift trigger and yet are comparable in cost. The MCR808 and the MCR1304 are interchangeable in all of the circuits shown in this section. All of these circuits are also suitable for other power control applications such as universal motor speed control which does not require feedback.

The circuit shown in Figure 10-2-1 is the simplest of the three versions. This circuit uses a switch, S_1 , to bypass the control circuit and thus provide full lamp brilliance. When S_1 is open and R_2 is set at minimum resistance, the lamp will be at about 70% of maximum brilliance.

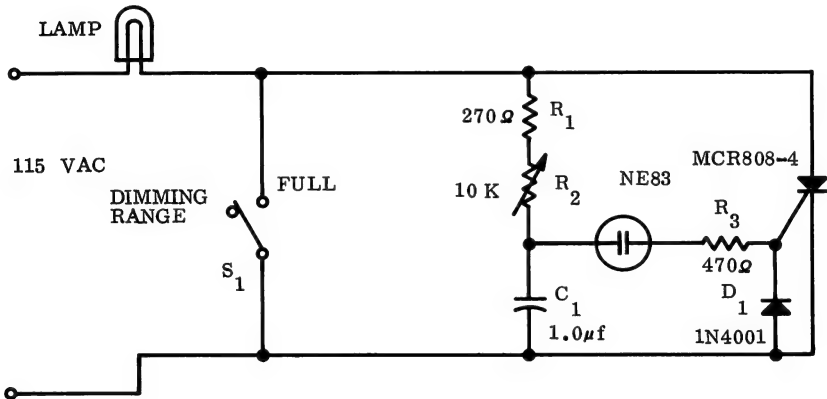


Figure 10-2-1 Half Wave Dimmer

By turning R_2 to maximum resistance, the lamp can be dimmed from the 70 level to zero. Operation of this circuit is rather simple. Capacitor C_1 charges thru R_2 and R_1 to the firing voltage of the neon bulb. Firing of the neon bulb triggers the SCR. The conduction angle of the SCR and, hence the brightness of the lamp are controlled by the setting of R_2 . R_3 and D_1 prevents gate avalanche by limiting the negative voltage applied to the gate.

The only difference between the circuit shown in Figure 10-2-2 from that just described is that it provides control over the range from full brilliance to zero. This is accomplished by switching out a rectifier and half of the potentiometer resistance at the midpoint of the dimming range. Sequence of operation is as follows: With the control set at full brilliance (minimum resistance), rectifier D_2 conducts for one-half of each cycle and

the SCR conducts for the other half cycle. As the potentiometer resistance is increased, the SCR conducts for an increasingly smaller portion of its half of the cycle. SCR conduction should reach zero at approximately the center tap of the pot. At this point, only the rectifier is conducting, supplying the lamp with half-wave power. S_1 is then actuated, switching out the rectifier and at the same time, shorting out the upper half of the potentiometer. Now, only the SCR is supplying the lamp with a half-wave power. As the control is rotated further, the potentiometer resistance increases once more, reducing the conduction angle of the SCR until it reaches zero conduction and zero lamp brilliance. The firing circuit functions as previously described.

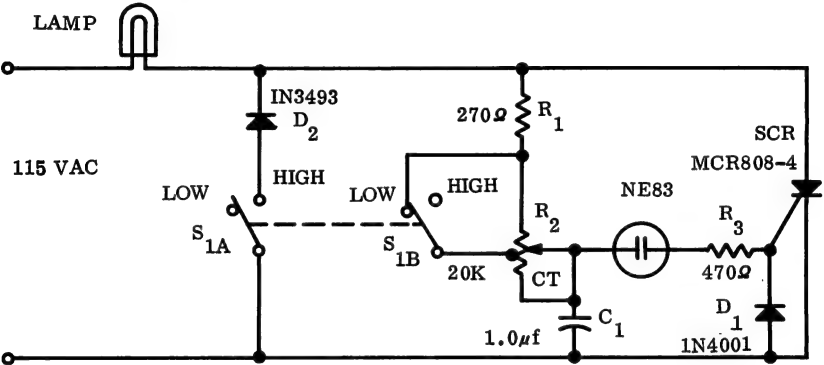


Figure 10-2-2 Center-Tapped Control Dimmer

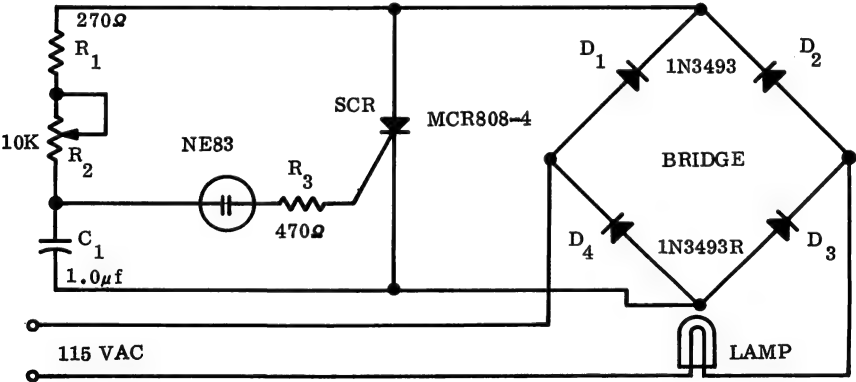


Figure 10-2-3 Full Wave Bridge Dimmer

The circuit of Figure 10-2-3 is advantageous because it provides a range from full brilliance to zero continuously without a switch actuation at the center of the range. This is accomplished by using a full-wave bridge which permits a single SCR to be phase controlled over both halves of the

cycle. Note that in this circuit, the clamping diode for the gate is not necessary since the control circuit is never reverse biased.

Full wave dimmer circuits giving continuous control have, in the past, been constructed using a single SCR across the dc terminals of a rectifier bridge. That approach was taken because the cost of a second SCR was greater than that of four rectifiers necessary to construct the bridge. The picture has been changed with the introduction of the Motorola MCR 1304. It is now more economical to use two SCR's for full-wave control, than one and a rectifier bridge. In addition, there are performance advantages for the two SCR circuit. In the rectifier bridge-single SCR circuit, the load current must pass through the SCR and two rectifiers during each half cycle, whereas in the two SCR circuit, the load current is carried only by a single SCR during each half cycle. Therefore, the power dissipation in the two SCR circuit is about 1/3 of that in the single SCR rectifier bridge circuit. The heat sink requirements are reduced proportionally. In the single SCR circuit, the SCR conducts all of the load current, whereas in the two SCR circuit, the current is divided. Therefore, for a given mounting base temperature, the RMS current carrying capability of the two SCR circuit is 1.4 times that of the single SCR circuit.

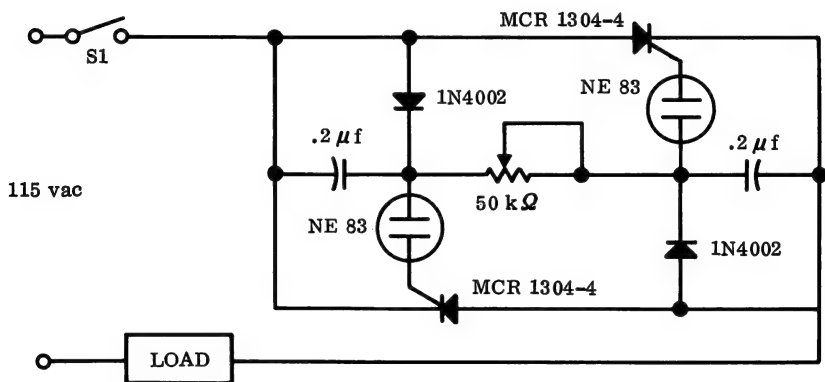


Figure 10-2-4 Two SCR Dimmer

Figure 10-2-4 shows a two SCR dimmer using a minimum number of components. In order to maintain low-cost, and at the same time provide reasonable control stability, neon bulbs were used as trigger devices. The NE-83 is a neon bulb which was designed for SCR triggering. It contains a trace of radioactive material to prevent its firing voltage from rising when it is operated in the dark.

A working model was constructed in a standard 2" x 2" x 4" steel electrical switch box. Each SCR was soldered to a 1-1/2" square x 1/16" thick copper heat sink. Electrical connections to the dimmer were made directly to the heat sinks to aid in heat removal. With a solid metal front plate and a single pair of No. 14 plastic insulated solid copper wire used for power connection, the heat sinks were found to have a thermal resistance

of approximately 12°C per watt to ambient. Under these conditions, the dimmer would be capable of handling a 300-watt lamp load at ambient temperatures up to 105°F . With a louvered switch plate or with more or larger conductors connected to the heat sinks, the unit would be capable of handling larger loads. The maximum limit would be about 11 amperes at a case temperature of about 70°C .

SECTION 11

Solid State Ignition Systems

The conventional ignition system is simple. There is a set of breaker points which interrupts an electrical current flowing through the ignition coil in exact synchronism with the travel of the pistons. At precisely the right moment, the breaker points open; the interruption of current induces a high voltage in the coil; the voltage is applied to a spark plug causing an electrical spark to ignite the gas in the cylinder; the gas explodes and the piston is started on another revolution. It's all very simple -- and it works. But every motorist soon becomes acquainted with the symptoms that spell TUNE-UP, which are caused by some of the inherent faults of conventional ignition systems. Many of these inherent faults can be eliminated or greatly reduced through the use of simple solid state ignition circuits.

To determine the reliable performance of any solid state ignition circuit, a knowledge of the advantages as well as the limitations of various types of applicable semiconductor products is required. The following discussion will describe some of the operating considerations in using presently available semiconductors, including some recently introduced new devices, in electronic ignition circuitry.

TRANSISTOR IGNITION CIRCUITS

The basic transistor ignition coil switching circuit is illustrated in Figure 11-1. The transistor is required to switch the coil current from a current level of 7-12 amp when turned on, to a voltage level from 60 - 120 volts when turned off. The zener diode protects the transistor against high voltage that could cause transistor failure. The capacitor across the coil performs the dual function of improving the switching load line of the transistor and tuning the coil for maximum voltage, especially with a low impedance loading on the coil secondary. The ignition coil and ballast resistor are designed for maximum ignition performance while maintaining semiconductor stresses within the safe and reliable region. The resistor in series with the base of the transistor must have a low enough value to insure low collector-emitter saturation voltage at every possible battery voltage condition when the transistor is turned on. The resistor connected between the base and emitter, insures a low collector-emitter current when turned off.

The circuit in Figure 11-2 is similar to that of Figure 11-1 except that the ignition coil is connected in series with the emitter of the transistor. In this circuit, the voltage across the primary of the ignition coil also appears across the distributor points. The secondary output of this circuit can suffer from the possible energy dissipated by arcing across the distributor points at low cranking speeds and moist atmospheres. When the coil is in the collector circuit, the breaker points only interrupt battery voltage.

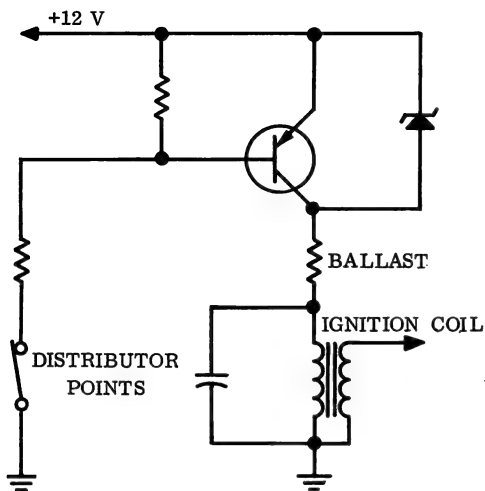


Figure 11-1 Basic Ignition Circuit
With Coil in Collector Circuit

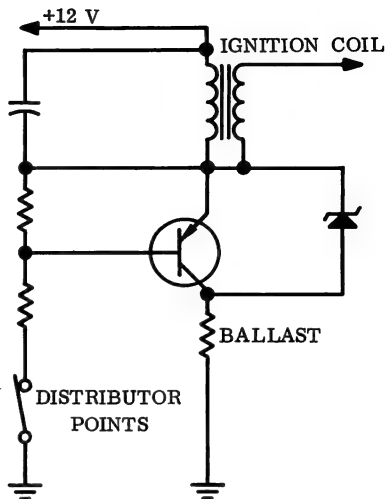


Figure 11-2 Basic Ignition Circuit
With Coil in Emitter Circuit

There are generally two basic ignition system designs. Both designs attempt to achieve the same high output voltage in the secondary of the coil. One design uses a coil with a turn ratio between a 230 to 325-1 which means that the secondary has 25,000 to 30,000 turns with a fairly fast rise voltage that can develop a high voltage across a fouled plug within the energy storage limits of a practical coil. The other system uses a high turn ratio coil of 325 to 500-1, with approximately the same number of primary turns, to develop a slower rise output that will generally result in a lower voltage across a fouled plug. Both systems have their merit, however, semiconductor requirements for each coil can be much different.

In selecting a transistor to operate in a particular ignition system, it is necessary to know the coil design and the voltage and current stress. A typical low turns ratio coil design will require between 7 and 12 amperes of collector current and a transistor with the ability to switch a load line up to 90 - 120 volts. This requirement is generally met by the graded base high voltage transistor such as the 2N2528 (Figure 11-3), or two lower voltage transistors in series as shown in Figure 11-4. The use of two fast switching transistors in series, to meet high voltage demands, is not recommended because of the greater difference in load line stress, due to the greater percentage difference in storage and switching time.

The high turns ratio coil will typically require a transistor that is able to switch a 7 -12 ampere collector current out to 65-90 volts. These requirements are generally met by a high gain alloy transistor such as the 2N1543 (Figure 11-5). This type of circuit can be protected by a zener diode connected from collector-to-base.

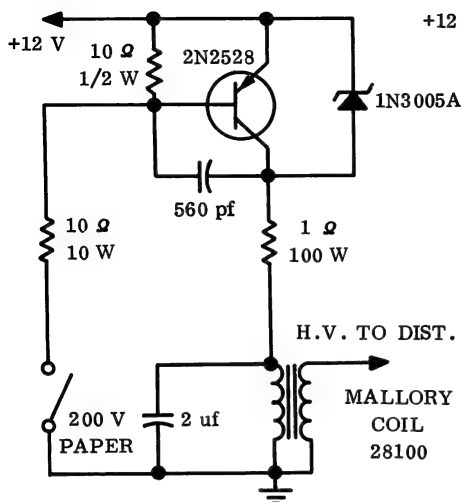


Figure 11-3 Ignition System Using Single Diffused Base Transistor

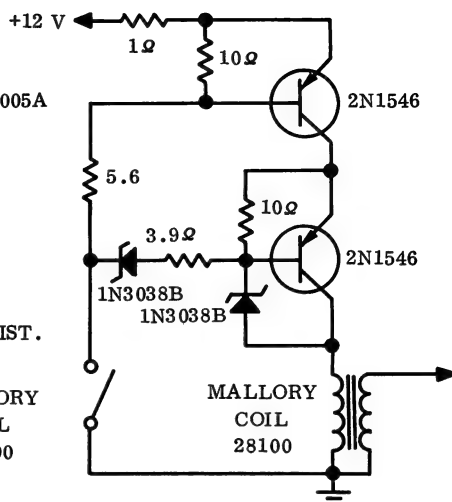


Figure 11-4 Ignition System Using Two Transistors in Series

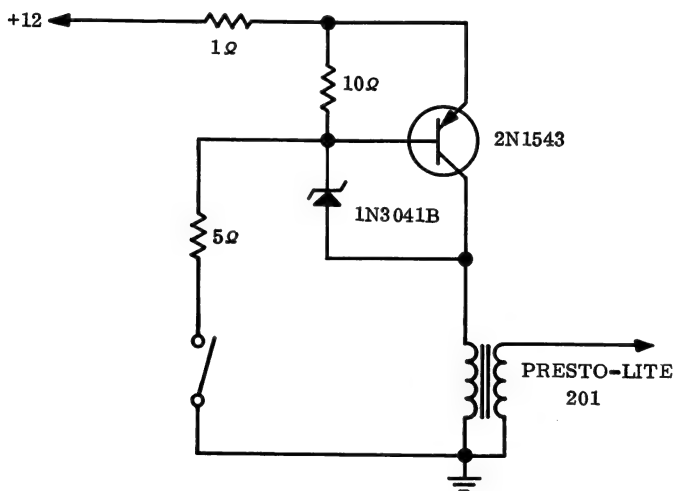


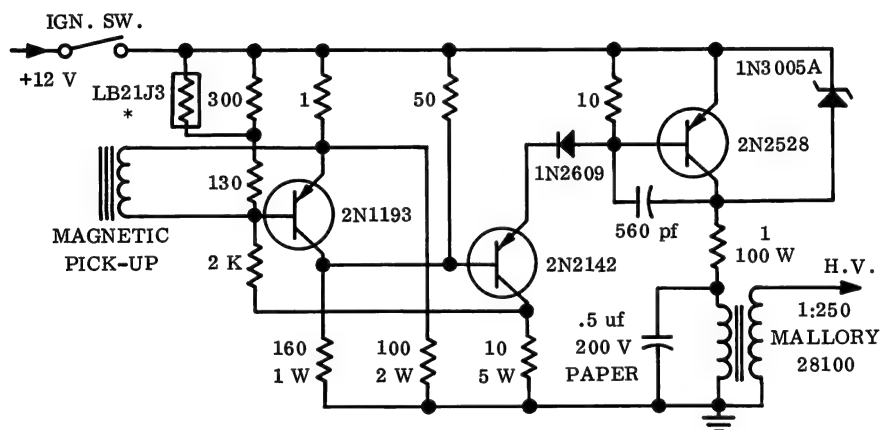
Figure 11-5 Ignition System Using Single Alloy Transistor

It is possible to connect three or more transistors in series to operate with the 3-4.5 ampere standard coil. This type circuit will generally limit coil voltage rise by avalanching and the transistor must be checked for these stresses.

CONTACTLESS IGNITION

A major improvement in ignition system performance and life can be realized when the mechanical distributor breaker points are eliminated. Basically, this involves a requirement for sensing the engine rotation and supplying a signal that is timed with the engine to switch the ignition coil circuit. The two common methods of non-mechanical rotation sensing are magnetic and photoelectric.

Figure 11-6 shows a typical contactless transistorized ignition circuit operating from a magnetic-pickup. The circuit is basically a bistable triggered flip-flop with sufficient gain to drive the ignition coil switching transistor. The point at which the coil circuit is triggered "on" and "off" is determined by the rotating teeth on the distributor shaft. A thermistor is inserted in the bias circuit of the 2N1193 to assure operation of the circuit at low temperature where distributor rpm is low and the pickup trigger signals are, in turn, very low. A silicon diode is inserted in series with the emitter of the 2N2142 driver transistor to provide bias for low "off" current in the drive circuit at high temperatures.



* Thermistor - Fenwal Electronics, Inc.
All Resistors 1/2 W Unless Otherwise Specified.

Figure 11-6 Magnetic Pulse Trigger Ignition Circuit

A contactless ignition system using a gate controlled switch is described later in this section.

TRANSISTOR LIMITATIONS (SAFE AREA)

To determine if the transistor is operating in the safe area region of the device, it is necessary to examine the operating load line of the ignition transistor. This display is observed on an X-Y oscilloscope

by connecting a collector current sensing resistor to the Y-axis (vertical input) of the scope and the collector-to-emitter voltage waveform to the X-axis (horizontal input). The Tektronix Model 536 is an ideal scope to display transistor load lines.

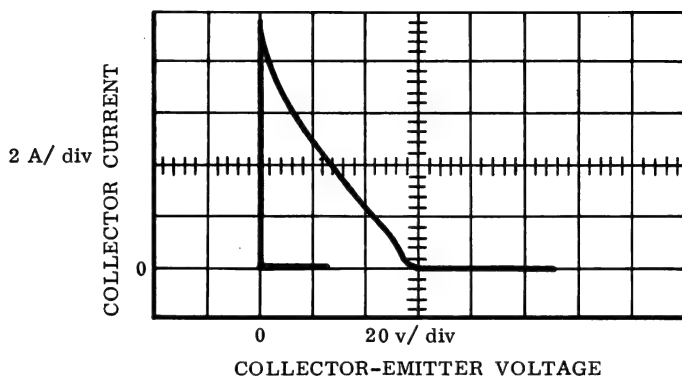


Figure 11-7 Load Line of Single-Transistor Circuit Shown in Figure 11-3

Figure 11-7 shows a typical load line display of the single transistor ignition system, shown in Figure 11-3 with the zener diode voltage limiting at 110 volts. By measuring the switching speed of the transistor in the ignition circuit, and correlating this with the transistor switching load line and "safe area" curve, as in Figure 11-8, the suitability of the transistor for the ignition circuit can be determined. It is particularly important to shield the scope leads when displaying the load line in the presence of the energy in the spark gap, and use a low inductance current sense device in order to eliminate distorted representations.

The "safe area" graph shown in Figure 11-8 was determined for the 2N2528 by stressing a quantity of transistors at a given voltage level with a timed current pulse. To prepare such graphs, the current levels and pulse widths that will produce destructive second breakdown of the transistors have been recorded and plotted as shown in Figure 11-9, the results are then translated to provide safe operating data (including a nominal safety factor) from the standpoint of voltage and current as a function of pulse length or switching speed.

The switching load line of the transistor is primarily dependent on coil design and on the switching speed of the transistor. Because of their basic design, most ignition coils have a considerable leakage inductance. This leakage inductance appears as an inductance in series with the coil primary, and a fast switching transistor will switch an inductive or voltage leading load line that can cause a voltage current excursion into an unsafe operating region.

By adding a capacitor in parallel with the primary of the ignition coil, the transistor will switch a capacitive or current leading load line

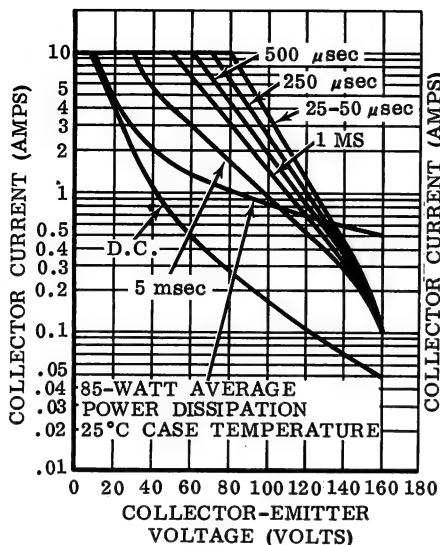


Figure 11-8 Safe Area For 2N2528

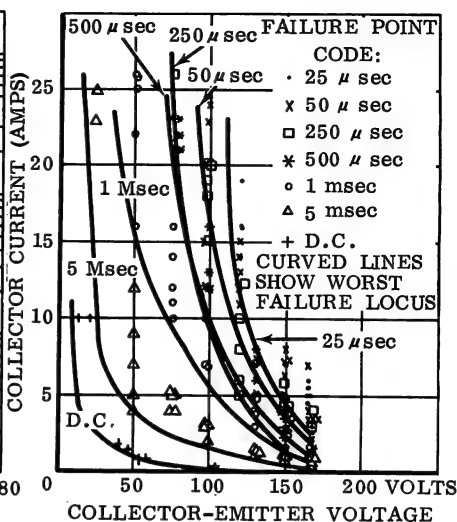


Figure 11-9 Failure Points For 2N2528

that can assure reliable operation. In addition, less switching power will be dissipated in the transistor. The larger the capacitor, the better the load line will appear. However, the coil output voltage can be reduced by this additional capacitive loading on the coil.

Sometimes the load line of an ignition coil switching transistor will extend into the third quadrant. The load line excursion into the third quadrant is caused by the ringing voltage of the secondary coil circuit. The voltage level in the third quadrant is generally clamped by the forward voltage drop of a zener diode connected between collector and emitter of the transistor. The current level is determined by the inverse gain of the transistor and the total external resistance path between base and emitter. If there is no forward diode connected between the collector and emitter of the transistor, the reliable operation in the third quadrant is determined by the voltage rating of the emitter-base junction and the voltage clamping action of the external resistance connected between emitter and base.

Load line excursion into the fourth quadrant is generally caused by pickup or small inductance in the sensing resistor, especially at fast switching speeds.

ZENER DIODE PROTECTION

The zener diode is an important component to assure solid state ignition system reliability. There are two basic methods of using a zener diode to protect an ignition transistor. These two methods are shown in Figures 11-3 and 11-5. In Figure 11-5, the transistor is protected by a zener

diode connected between base and collector and in Figure 11-3, the zener is connected between emitter and collector. In both cases the voltage level of the zener must be selected carefully so that the voltage stress on the transistor is in a region where the safe operating area is adequate for reliable circuit operation.

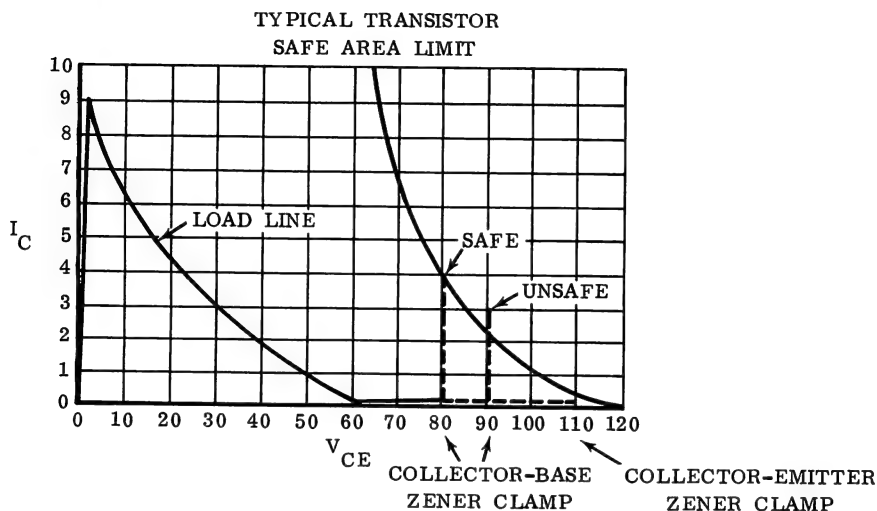


Figure 11-10 Safe Zener Protection

Figure 11-10 illustrates "safe" and "unsafe" selection of a zener diode for collector-base protection of a transistor in the same ignition coil circuit. It can be seen that the safe operating area of a transistor must be known if an adequate protective zener is to be selected. Additionally, it can be seen that collector-emitter protection of the transistor can allow higher voltage stress of the transistor.

The zener diode must be able to take the stress of peak pulse current necessary to clamp the voltage rise across the transistor to a safe value. In a typical case, a 10-watt, 100-volt zener diode is required to operate with an 80- μ second peak pulse current of 8 amperes when connected between the collector-emitter of the transistor. The waveform of this pulse current approaches a sine wave in shape (Figure 11-11). The voltage rise across a typical 10-watt zener diode due to this current pulse is shown in Figure 11-12. This voltage rise of approximately 8 volts indicates an effective zener impedance of approximately 1 ohm. However, a good share of this voltage rise is due to the temperature coefficient and thermal time constant of the zener. The temperature rise of the zener diode junction is indicated by the voltage difference between the rise and fall of the current pulse.

It must be remembered that safe operation under this peak pulse current is not tested by the manufacturer unless a peak pulse test is specified, and a possible failure with "off-the-shelf" zeners can be the result of this stress. It is possible that voltage excursions above 10 volts

can be encountered under this 8-ampere, 80- μ second pulse current.

When the zener diode is connected between the collector and emitter of the transistor, an additional power dissipation will result from the clipping of the ringing voltage of the ignition coil by the forward conduction of the zener diode. This power dissipation by the forward diode current will result in additional zener voltage rise depending upon the temperature coefficient of the device. It is not uncommon to observe a 15-volt rise above the zener device voltage rating due to temperature coefficient and impedance under these pulse current conditions.

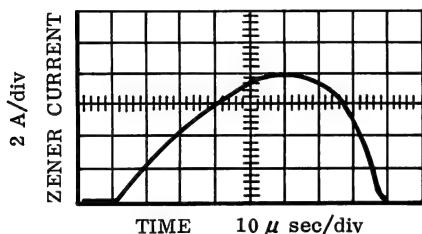


Figure 11-11 Zener Diode Current Pulse

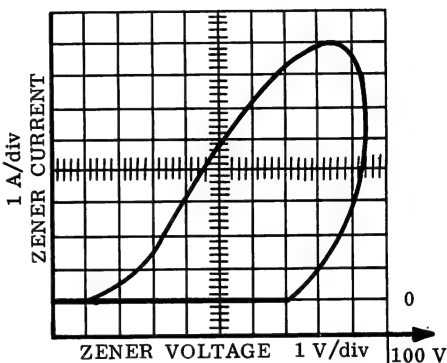


Figure 11-12 Voltage-Current Representation on 100 V Zener

In many cases, a 1-watt zener diode will be sufficient to protect the transistor. In some 1-watt zeners, the junction area is the same as that of 10-watt zener diodes. The junction-to-case thermal resistance is higher due to the steel case, but additional power can be dissipated by heat sinking the case with a fuse clip thermally connected to the circuit package heat sink. The reliability of the 1-watt unit under high pulse currents is slightly reduced from that of the 10-watt device, but the voltage rise will be greater under the same pulse current conditions due to higher junction temperatures and differences in thermal time constants.

The zener diode should be connected as close as possible to the terminals of the transistor to be protected. This insures that induced voltage transients, caused by current changes in long lead lengths, are clamped by the zener and do not appear across the transistor.

IGNITION SYSTEM FAILURE POSSIBILITIES

In some cases, transistor failures have occurred even after a cautious approach has been taken to insure reliable load line operation. There are several possible reasons for an ignition transistor failure that are not generally taken into account when designing an ignition system.

With reference to the basic transistor ignition system (Figure 11-1), it can be shown that if the distributor points were closed, approximately 60 - 100 μ seconds after opening, the transistor would have to turn on from a voltage level equal to the zener diode voltage level. This load line (Figure 11-13), will initially switch out in current to the zener diode pulse current level, and then may follow a capacitive load line back to the saturation voltage of the transistor. This can produce a greater stress across the transistor than the load line caused by switching off the current level in the ignition coil. Any distributor point bounce at the break would have to be coincidental with the point of maximum zener pulse current in order to produce maximum stress of the transistor.

This same load line stress can occur if the battery connection is intermittent while the distributor points are closed. Of course, any secondary coil load condition that would reduce primary voltage swing or any point bounce action before or after maximum voltage rise, would reduce this load line stress. Ignition system operation with circuitry to eliminate the use of breaker points will eliminate the problem caused by bouncing distributor points, but will not eliminate the problem caused by the intermittent ignition key action.

Referring again to Figure 11-1, it can be shown that when the ignition switch is opened (points are closed), the voltage rise across the coil can cause collector-base failure or base-to-emitter failure of the transistor. The voltage stress on each of these junctions depends upon the zener diode current that flows through the resistor in series with the base, and the resistor connected between base and emitter. These junctions can be protected by connecting a diode between base and emitter, with the cathode of the diode connected to the emitter.

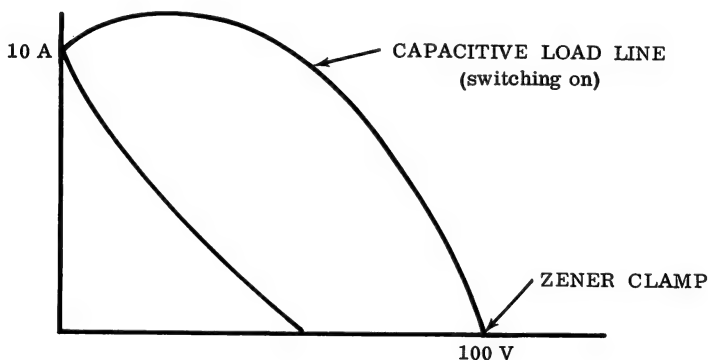


Figure 11-13 Switching "On" From Zener Level in Circuit of Figure 11-3

It has been noted that "pickup" on the lead connected from the distributor to the base of the transistor may cause transistor failure. These failures were especially prevalent when large sparks or cross fire were purposely produced in the distributor cap. The addition of a small capacitor connected between the base and collector, or a larger

capacitor connected between the base and emitter, can greatly reduce this failure problem.

Whatever is determined to be the possible cause of transistor failures using transistors with high "safe area" at the zener diode voltage level will tend to eliminate the failure mode.

REVERSE BIAS CIRCUITS

Ignition transistor switching losses can be greatly decreased and ignition coil output increased by reverse biasing the transistor to decrease switch-off time. The reason for decreased switching time is the ability of the reverse bias circuit to drive out the charge stored in the base region of the transistor more rapidly.

A basic circuit to decrease the switching time in a transistor is shown in Figure 11-14. The charge on the capacitor, obtained during the time when Q_1 is not conducting, supplies the reverse bias for the coil switching transistor when Q_1 is triggered "on". The driving pulse can be supplied by a monostable multivibrator that may be a part of a magnetic pulse triggered amplifier circuit. The ignition transistor is normally "on" and is triggered "off" for a time long enough to allow the energy stored in the coil to dissipate through the spark plug. The length of time that current will flow through the spark plug can be limited by the length of the trigger pulse. This can reduce electrical gap erosion in the spark plug.

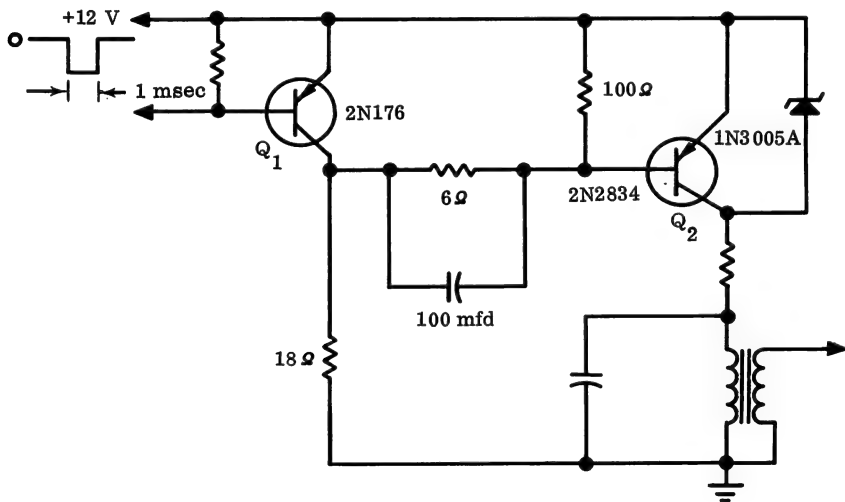


Figure 11-14 Capacitor-Discharge Turnoff Circuit

The main consideration in any reverse bias circuit is the breakdown voltage of the transistor emitter-base diode. Most high voltage, graded base transistors have a low emitter-base breakdown voltage.

When using a 2N2834 transistor, maximum reverse voltage should be limited to 3 volts. The 2N2528 transistor can safely operate with reverse bias voltage up to 6 volts.

Figure 11-15 shows several additional methods of providing reverse bias for the transistor. These circuits use diodes, transformers or inductors to provide a reverse bias voltage. The inductors or transformers develop a bias voltage that is induced by either a change in the current level in the base or in the collector circuits of the transistor. An important point to remember here is that this phasing of feedback voltage can cause the circuit to oscillate if the feedback loop gain is high.

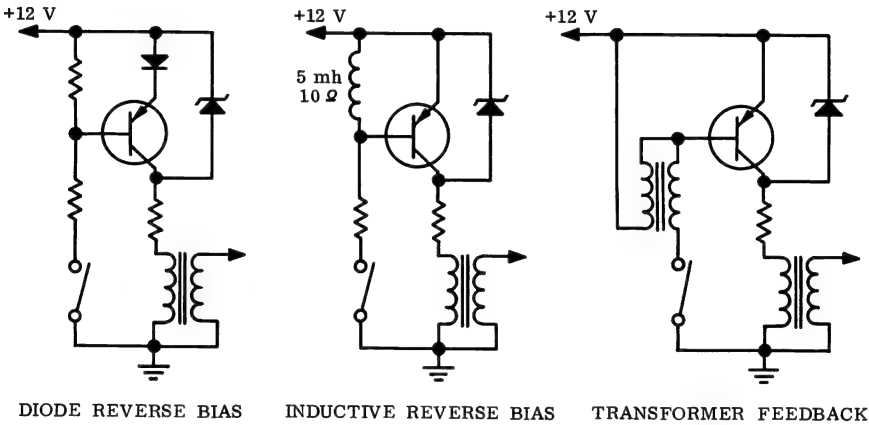


Figure 11-15 Reverse Bias Circuits

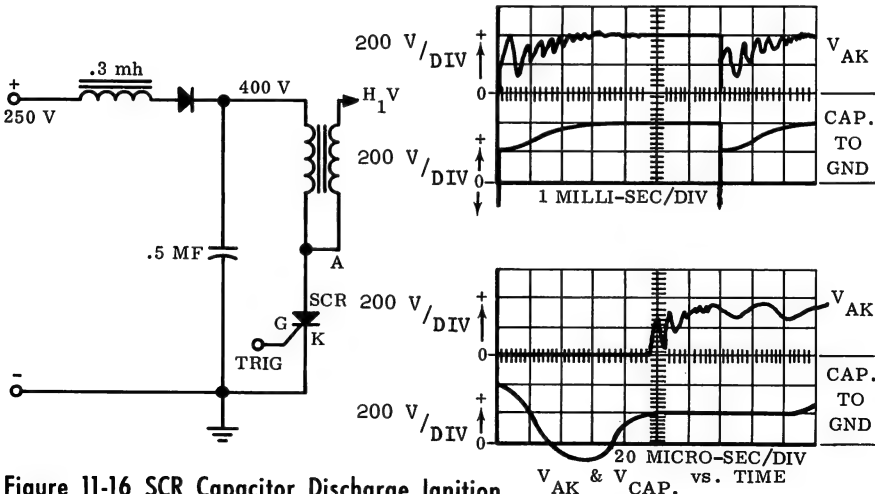


Figure 11-16 SCR Capacitor Discharge Ignition
(Circuit and Waveshapes)

SCR CAPACITOR DISCHARGE IGNITION

An alternate system of producing the high voltage necessary for automotive ignition, is the capacitor discharge ignition system. In this system a capacitor is charged to a high voltage level and then abruptly discharged through the primary of an ignition coil as shown in Figure 11-16. The silicon controlled rectifier (SCR) performs the function of discharging the capacitor when triggered in synchronism with the spark requirement.

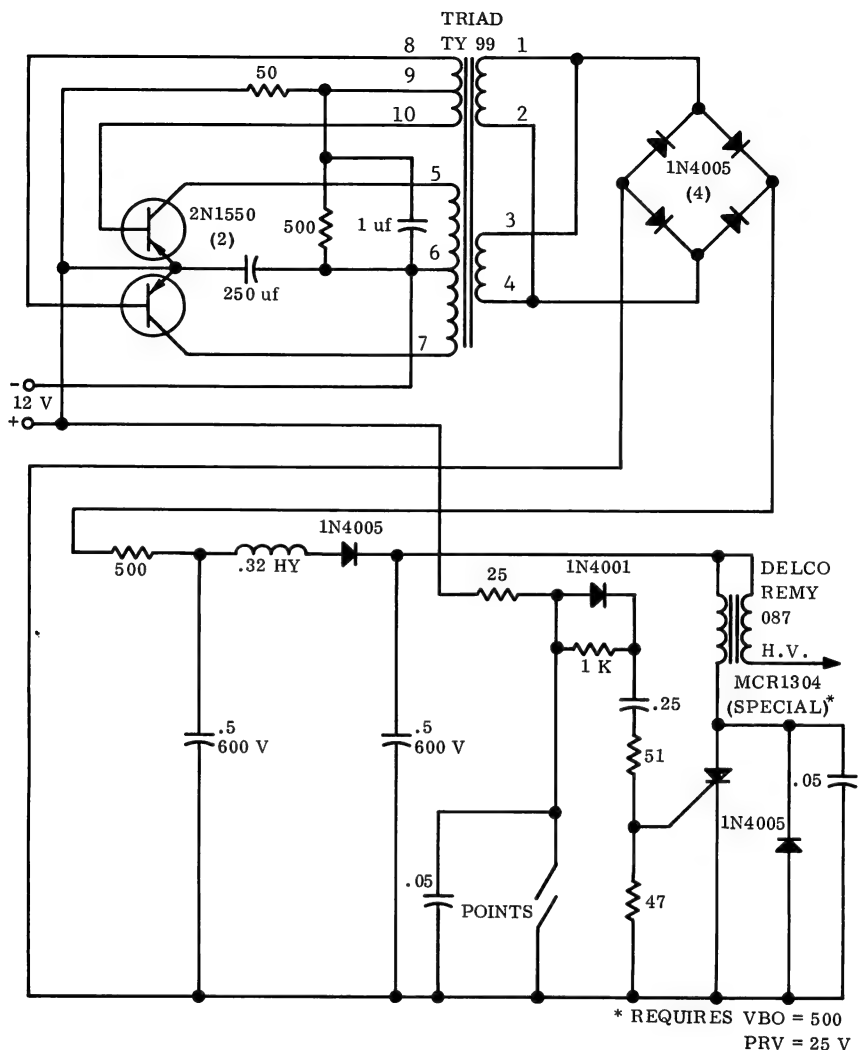
The SCR is a four layer semiconductor device that will block a high voltage until a trigger pulse is applied between the gate and cathode. When this gate signal is applied, the SCR will turn "on", current will flow through anode and cathode and will continue to flow until forward voltage is interrupted or voltage is reversed. The SCR will turn "off" in the capacitor discharge system (Figure 11-16), by the reversal of voltage due to the oscillating action of the coil and capacitor.

Because the energy that can be stored in the capacitor is proportional to the square of the voltage, it becomes more economical to use high voltage as far as the capacitor is concerned. However, the higher the voltage that the capacitor is charged to, the higher the cost of the SCR. The practical voltage level of operation would be 200-400 volts. The design of the ignition coil will also determine the design voltage level.

The 200-400 volts can be obtained from 115 volts ac, in stationary engine installations, or from a one or two transistor dc to dc converter circuit operating from the vehicle battery. A two transistor converter circuit is shown in the capacitor discharge circuit of Figure 11-17.

A single transistor triggered blocking oscillator voltage supply for a capacitor discharge ignition is illustrated in Figure 11-18. In this circuit, the transistor is triggered "on" for a single cycle by the closing of the points. The capacitor is charged by the transformer output pulse and restricted from discharging by the silicon diode. When the breaker points are opened, the SCR is provided with a trigger pulse which discharges the capacitor through the ignition coil and supplies ignition voltage to the spark plug. When the points again close, the cycle will repeat. This circuit can be easily adapted to contactless ignition without breaker points.

In the two transistor and 115 volt ac source circuits for the dc voltage, resonant charging of the capacitor is recommended. This charging circuit can theoretically charge the capacitor to twice the input voltage level at all engine speeds, however, 1.5 times is a more practical value. Operation with the series inductance and without the diode can charge the capacitor to peak voltage levels of 1.5 times the supply, but the capacitor will discharge from various voltage levels, depending on engine speed. The series inductance should be high enough to present a high impedance to the power supply while the SCR is conducting and low enough to allow the capacitor to fully charge before the SCR is triggered. Between 100-300 millihenry inductance is a good practical range for this type ignition system design.



**Figure 11-17 Two Transistor Inverter For
SCR Capacitor Discharge Ignition**

One of the main advantages of capacitor discharge ignition is the design availability of fast rise time ignition voltage. These fast rise times can be achieved economically by discharging high peak currents through low inductance coil designs. Capacitor discharge ignition coils can be designed with voltage rise times of less than $5 \mu\text{seconds}$ that can produce adequate ignition voltages across a very low resistance fouled plug condition.

TABLE 11-1

Circuit	Secondary Voltage			Ave. Bat. Current	
	1000 rpm		5500 rpm	1000rpm	5500rpm
	1 Meg.	50 μ f	50 μ f		
Blkg. Osc. Cap. Disc. Prestolite Coil Figure 11-18	19.5 KV	27. KV	27. KV	800 MA	2.8 A
One Trans. Ignition Prestolite Coil Figure 11-5	14.5 KV	27. KV	23.5 KV	5.2 A	2.4 A
Two Trans. Cap. Disc. Delco Remy Coil Figure 11-17	23 KV	29. KV	29. KV	930 MA	1.75 A
Standard Ignition Delco Remy Coil	20 KV	33. KV	20. KV	1.9 A	950 MA

Several important considerations in designing a SCR capacitor discharge ignition system are as follows:

1. The gate trigger pulse time should be at least 20 μ seconds shorter than the 1 cycle time of the resonant frequency of the coil. This will assure that the SCR will be reversed biased for turn-off. In the circuit of Figure 11-18, the time after capacitor discharge and before reapplication of voltage is relatively long and SCR trigger pulse time is not critical.
2. The maximum resonant frequency design of the coil is limited by the turn-off time of the SCR. This maximum frequency will be about 25 to 30 KC for presently available low cost SCR's.
3. When fast high current capacitor discharge circuitry is desired, the $\frac{di}{dt}$ characteristics become important. This characteristic of the SCR will offer a highly varying series resistance to the initial flow of current for a limited time which can greatly reduce the first cycle of the output voltage.
4. A small capacitor should be connected in parallel with the SCR so that reapplication of voltage is slowed below the $\frac{dv}{dt}$ rating of the SCR. This is especially important in using high leakage inductance coil designs. This precaution is necessary to prevent the SCR from latching "on".
5. It sometimes becomes more economical to parallel the SCR with a diode to eliminate a requirement for reverse voltage rating.

GATE CONTROL SWITCH IGNITION

One of the most recent semiconductors to become available for electronic ignition systems is the silicon gate control switch (GCS). The "turn on" action of this semiconductor switch is like the SCR. However, it is no longer necessary to interrupt the supply voltage or reverse bias the anode cathode circuit to turn it off. The load current can be turned off with a relatively small reverse voltage pulse applied at the gate. (See Section 2.)

Because of its ability to block high voltage levels and only pulse power is necessary to switch from "on" to "off", this device becomes very desirable for ignition application. The present available product for ignition has the ability to turn off a 7-ampere coil current to voltage levels approaching 500 volts at 100°C. The characteristics of the GCS can allow ignition coil designs to be very similar to conventional designs.

A basic silicon gate controlled switch circuit is illustrated in Figure 11-19. To operate with a conventional distributor, the GCS must be turned on when a voltage is induced in the secondary of the trigger transformer, due to primary current flow caused by distributor point closure. The positive voltage of the trigger transformer secondary appears at the gate of the GCS, triggering the device and causing current to flow through the ignition coil. Energy is stored in the trigger transformer while the points are closed and the GCS is conducting. When the points open, this energy induces a voltage of the opposite polarity to turn the GCS off.

A simple circuit for storing the energy necessary to turn off the GCS is shown in Figure 11-20. When the breaker points close, the transistor is turned off and the charging current through the capacitor will turn on the GCS. The capacitor will charge to a voltage level that is dependent upon the energy stored in the inductance while the transistor is on. In this case 25-30 volts.

When the points open, the transistor will turn on and discharge the capacitor through the gate circuit with a polarity that will turn the GCS off. The interruption of coil current will supply the required ignition voltage. A low voltage GCS can also perform the function of the transistor in the circuit.

The resistor connected to the gate of the GCS will perform the function of turning the GCS back on after the capacitor has discharged, even before the points have closed again. This allows more time for energy storage in the ignition coil with a resultant improvement in high speed ignition performance over the conventional system.

A useful GCS ignition system is shown in the blocking oscillator circuit of Figure 11-21. In this circuit a magnetic or photoelectric device can supply the trigger for the one-shot blocking oscillator. The feedback winding continues to turn the transistor on until the core saturates. During the time the transistor is "on" the GCS is supplied with a off drive pulse at the gate. When the core saturates and the transistor turns off, an on

drive pulse is supplied to the gate of the GCS. The transistor will remain off until the next trigger is supplied.

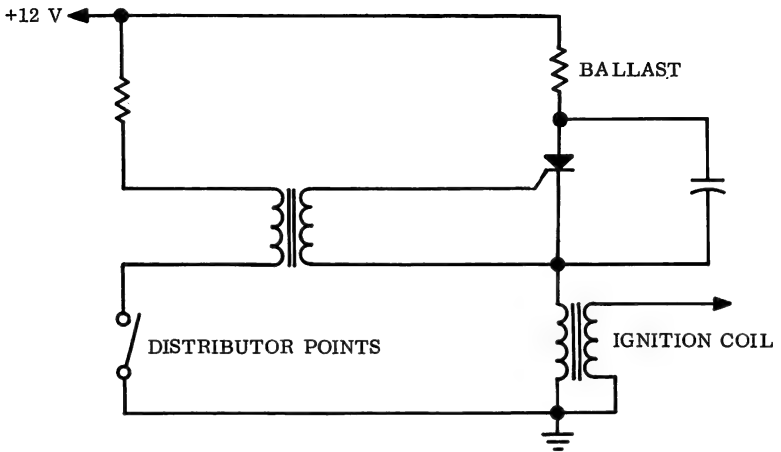


Figure 11-19 Basic Gate Controlled Switch Ignition Circuit

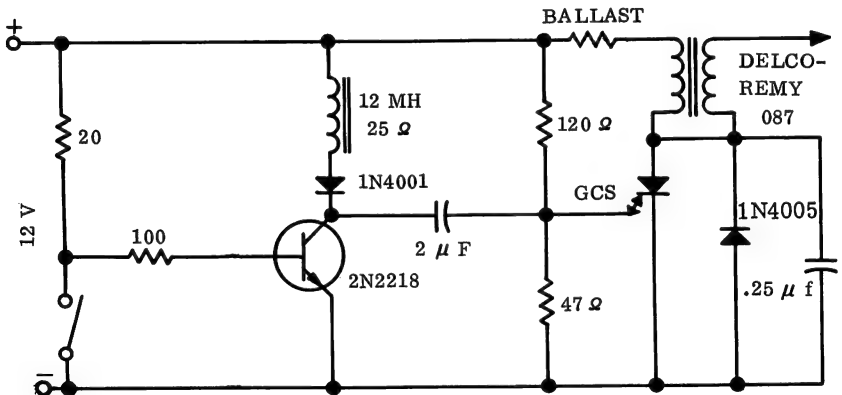


Figure 11-20 GCS Capacitor Turnoff Ignition

Some important considerations in utilizing the GCS in an ignition system are as follows:

1. A small capacitor should be connected between the anode and cathode of the GCS to prevent switching an inductive load line that could lead to device failure. This capacitor can also assist

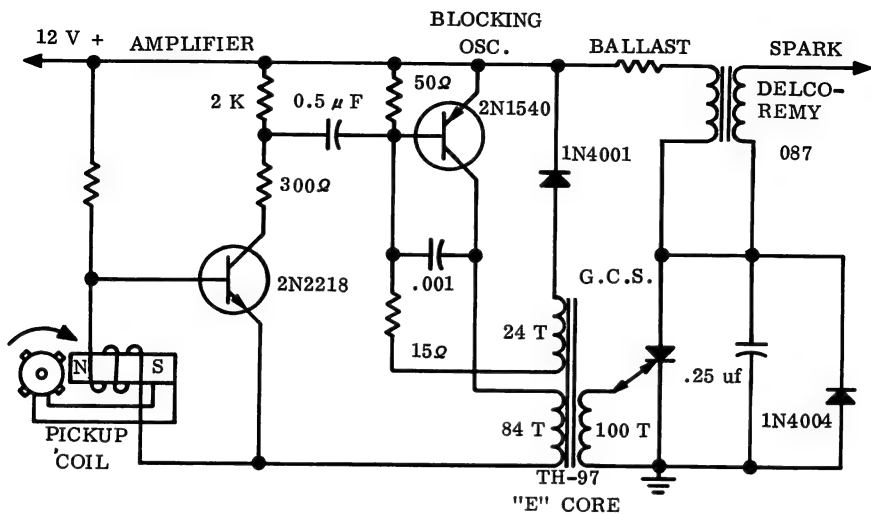


Figure 11-21 GCS Triggered Blocking Oscillator Ignition

the turnoff gain of the device by shunting some of the anode current during cathode to anode voltage rise.

2. A silicon diode connected between anode and cathode will eliminate the requirement for a reverse voltage rating for the GCS.
3. The GCS should not be turned off with borderline trigger pulses. It is possible to "hang up" at some high power dissipation point during the switching off process and thermally damage the device. Overdriving with the turn-off trigger will also decrease switching time. Peak off drive gate currents of 2 amperes for 20-50 μ seconds is recommended for anode current levels, 5 amperes or above.
4. At low temperatures it becomes much easier to turn off the GCS, but turn-on characteristics present some limitations. In order for the GCS to latch on, an anode current level must be reached where the product of the current dependent gains become greater than 1. At low temperature this latch on current level becomes quite high and the gate trigger pulse must be held for a length of time great enough for coil current to reach this level with the time constant of the ignition coil circuit.

SECTION 12

RF Transmitters

Introduction and General Considerations

The state of the art in transistorized transmitters is presently advancing rapidly in regard to power output and operating frequency. The prospect of generating watts of power at hundreds of megacycles directly from dc supply voltages of 12 to 24 volts is appealing to those designers of equipment in which space, efficiency and reliability are of prime concern. However, before attempting to reap the benefits of the technological advances made by the semiconductor industry in RF power, the following cautions and recommendations which pertain to RF devices in general, regardless of manufacturer should be considered.

CLASS C VS. CLASS A OPERATION

In order to obtain the unique device characteristics necessary for RF power generation, certain desirable characteristics normally found in power transistors are necessarily degraded. Probably the most outstanding example of this is dc operating safe area. This parameter is, to a large extent, dependent upon base width with a wider base providing an improved safe area. (Refer to Section 5 for a more detailed explanation of safe area.) Base width must be kept small (on the order of $1/2$ to $1-1/2$ microns) to achieve RF power gain. Therefore, any device designed for high frequency class "C" operation will be unsuitable for use in a dc biased class "A" circuit at even a fraction of its rated power output. In fact, in a circuit where bias is used only to overcome the base emitter offset voltage such as a class "B" linear stage, safe area problems can be encountered if too much idle current is applied. For example, the 2N2947 has a maximum continuous collector current rating of 1.5 amperes. A class "C" continuous current of one ampere would be considered safe when operating from a 25-volt supply. However, the one ampere of collector current is the average value of the pulses of collector current which flow only when the transistor is at or near saturation and the collector-emitter instantaneous voltage is low. If a continuous dc voltage and current of 25 volts and one ampere were applied at 25°C , the maximum device dissipation would not be exceeded. However, the device would be instantly destroyed due to exceeding the safe area. For similar reasons, RF power transistors should not be tested for collector to emitter breakdown on a curve tracer since the device is usually held in an unsafe area for a period of time that is essentially a dc test for purposes of defining safe area. A pulse technique should be used with a low duty cycle (2%), to avoid junction damage when testing for breakdown.

PULSED VS. CW POWER

When using transistors to generate high power at VHF, the factors which limit power output in most cases are other than power dissipation. High frequency transistors are usually peak voltage or peak current limited

before the dissipation limit is reached in an actual operating high frequency class "C" circuit. The peak voltage limitation can be exceeded by either excessive collector supply voltage, operation into an inductive load, or a combination of both. If insufficient current limiting is employed, exceeding the peak voltage rating it usually results in the destruction of the device. Attempts to use sufficient series resistance to limit current peaks during voltage breakdown will usually result in too great a sacrifice in power output during normal operation to make this system practical. Therefore, the peak voltage rating of a device is a limit which must not be exceeded in a practical circuit. Collector-to-emitter waveforms taken using a sampling oscilloscope will indicate if peak collector-emitter voltages are within ratings. The peak voltage in a circuit should be checked at the worst case conditions of detuning and unloading. Since the peak voltages are occurring at the carrier frequency, the same voltage limit applies for pulsed as well as C. W. operation.

A transistor behaves similar to a emission limited vacuum tube in regards to peak current handling capability. The amount of peak instantaneous current available is determined by the transistor geometry and no reserve or "space charge" effect exists. Therefore, if a transistor is operated with maximum collector-to-emitter voltage and drive is increased until there is no further increase in output, it can normally be assumed that the maximum peak current has been achieved. This effect is a gradual one since some additional power output will result due to the increased conduction angle of collector current with heavy base drive after the peak current capability of the device has been reached. This effect is particularly pronounced as the high frequency limit of operation for the device is approached.

From what has been stated here, it is logical to assume that relatively little more peak power is attainable from pulsed operation of a device than that which is available from CW operation. Some increase in peak pulse power will be noted over CW which can be attributed to the positive thermal coefficient of r_b' and collector bulk resistance. This increase in peak power might be as much as 20 percent in some cases but never as much as could be expected from a dissipation limited device as is often the case in vacuum tubes.

AMPLITUDE MODULATION

Transistors will collector modulate in a manner similar to that of vacuum tube plate modulation systems, with a few exceptions. Since transistors are normally current and voltage limited, a device which is already at the limit of collector supply voltage and current will not up-modulate from that power level. For a collector modulation system to function, a collector supply voltage of one-fourth the value of maximum rated voltage must be used. Since the peak current must double in addition to the voltage, a carrier level of one-fourth maximum power output must be maintained if 100 percent up-modulation is desired.

Feed-through capacitance in transistors will allow a residual carrier to be passed from the driver through the final even if the down-modulating audio has reduced collector to emitter voltage to zero. Therefore, some

modulation of the driver is required to achieve good down-modulation of the final. Moreover, up-modulation of the driver will aid in achieving the higher peak current required from the final on up-modulation. The effects of carrier feed-through can be minimized by the use of a common-base configuration in the final.

SECTION 12-1

Marine Band Applications

The marine band equipment is usually of the radio-telephone type with transmitter power ratings¹ varying from 15 watts to several hundred watts. The frequency range is generally 2 Mc to 3 Mc, with crystal controlled channels.

A 20-WATT 2.18 Mc TRANSMITTER

CIRCUIT DESCRIPTION

The 20-watt transmitter (see Figure 12-1-1) consists of an output, driver and oscillator stage. Both output and driver stages are operated common-emitter with the collectors grounded directly to the chassis. Collector-type modulation is applied to the emitter-base junction of both the driver and output stage. The supply voltage is 12 volts.

The output stage (2N2832)² delivers 10 watts (rms) carrier power into a 50 ohm load. The 50-ohm load simulates the approximate input impedance of a 9-foot bottom-loaded whip antenna. A capacitive-tap, pi-type network is utilized to transform the 50-ohm load to approximately 7 ohms at the emitter. The 7-ohm emitter loading is approximately the impedance the emitter must work into from a 12-volt supply to make the required 10-watt carrier power.

The required drive power to the output stage for rated power output is approximately 1 watt. The input impedance of the 2N2832 is approximately 5 ohms. This impedance level is transformed to the required driver loading impedance with a toroid transformer. This transformer is bifilar wound to reduce the leakage reactance. The remaining leakage reactance and primary reactance are tuned out on the primary side.

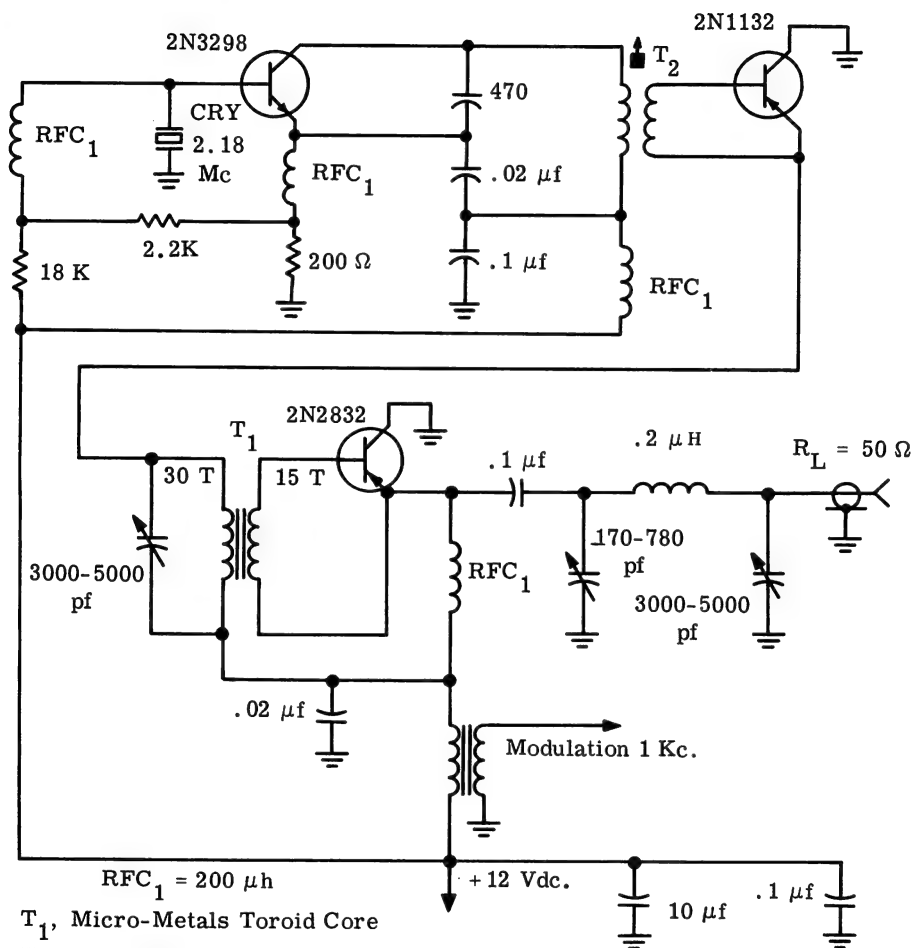
The driver stage (2N1132)³ provides approximately 1 watt carrier power (rms) to drive the output stage to rated power output. The required emitter loading for this power level is approximately 72 ohms.

The driver stage input network is also a toroid type transformer. This network provides the double-ended drive required to retain common-emitter operation in the driver circuit and also provides the collector coil of the oscillator. The required power level is approximately 40 mW.

-
- 1) The power rating is the dc power required by the transmitter final.
 - 2) The 2N2832 is an epitaxial base power transistor with an F_T of 10 Mc.
 - 3) The 2N1132 is a PNP silicon transistor with a maximum power handling capability at 2.18 Mc equal to approximately 2 watts.

PERFORMANCE

The transmitter delivers a 10-watt carrier (rms) into a 50-ohm load with an overall efficiency of 48 per cent. The output stage collector efficiency is 50 per cent. The required modulation power and resistance is approximately 7 watts and 10 ohms.



T_1 , Micro-Metals Toroid Core

T-50-2 Bifilar

T_2 , Pri. 40 T, #28, 1/4" ID Slug Tuned Form

Sec. 8 T, #28, Wound over Pri.

Figure 12-1-1 20-Watt 2.18 Transmitter

A 40-WATT 2.18 Mc TRANSMITTER

CIRCUIT DESCRIPTION

Transmitter two (see Figure 12-1-2) consists of an output and driver stage. Both stages are operated common emitter with the collectors grounded directly to the chassis. The modulation is applied in the same manner as described for the 20-watt transmitter.

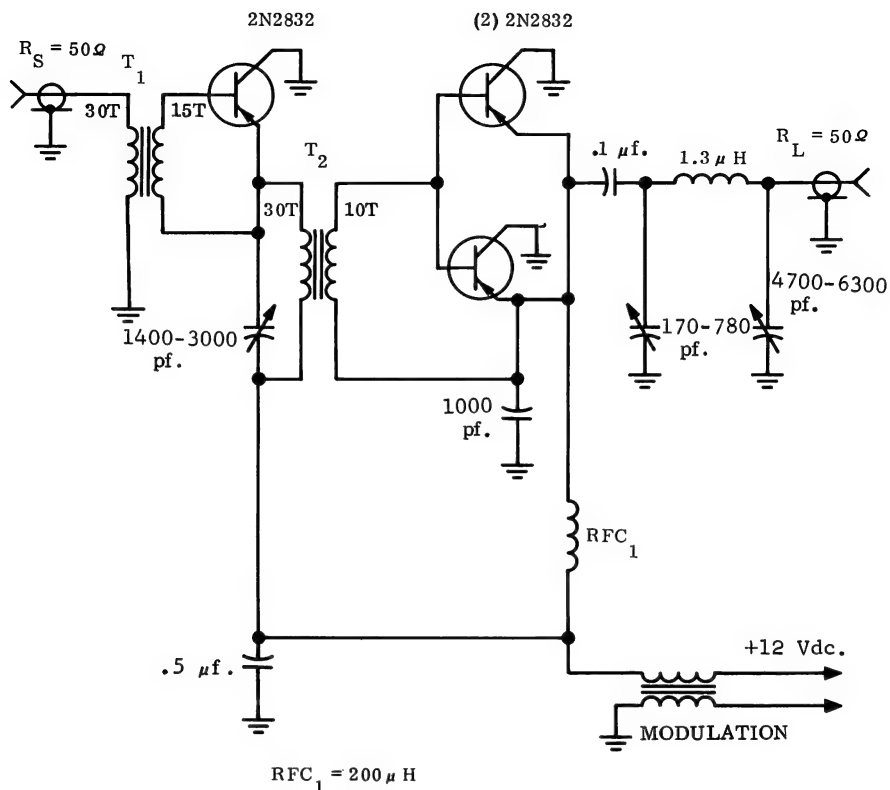
The output stage (two 2N2832's in parallel) delivers 20 watts (rms) into a 50 ohm load. A pi-type network is utilized to transform the 50-ohm load to approximately 3.5 ohms at the emitter.

The input impedance is approximately 3 ohms. This impedance is transformed to the required driver emitter loading by a toroid transformer. The required drive level for rated output power is approximately 2 watts. The correct driver loading for this power level is approximately 36 ohms.

The input network to the driver stage is also a toroid transformer. This transformer matches the input impedance of the driver stage to a 50-ohm driving source. The required input power is 25 mW.

PERFORMANCE

This transmitter delivers a 20-watt carrier (rms) into a 50-ohm load with an overall efficiency of 48 per cent. The output stage efficiency is 50 per cent. The maximum upward modulation is approximately 80 per cent. The required modulation power and resistance is approximately 18 watts and 5 ohms respectively.



$RFC_1 = 200 \mu H$
 T_1, T_2 , Micro-Metals Toroid Core T-50-2 Bifilar

Figure 12-1-2 40-Watt 2.18 Mc Transmitter

SECTION 12-2

A Citizen Band Transmitter

The 2N2950 transistor has made it possible to build citizens band transmitter final stages with a carrier power of two watts without resorting to paralleling of devices. Also, the stud mount and three completely isolated leads are features that solve mounting and heat sinking problems and allow complete freedom in the choice of circuit configuration. The 2N2951 is ideally suited as an oscillator and driver transistor and complements the 2N2950 in citizen band circuitry.

TRANSMITTER

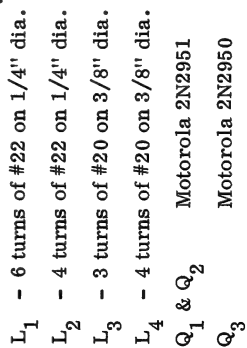
Figure 12-2-1 is the schematic drawing of a 2 - watt carrier citizens band transmitter. The performance data is shown in Table 1.

TABLE 1

Performance Data

	<u>Unmodulated</u>	<u>Modulated</u>
RF Power Output	2.0 W	2.6 W
DC Power, Final	3.05 W	3.05 W
Audio Input Power	---	1.5 W
Audio Modulating Impedance	---	39 ohms
Total Transmitter Current	360 mA	330 mA
Final Current	255 mA	230 mA *
Modulation Distortion	---	Less than 5%
Percent Modulation (upward)	---	80% Min.

* Caused by driver modulation



SECTION 12-3

A 30 MC SSB Transmitter

Figure 12-3-1 is a preliminary circuit showing suggested circuitry for using Motorola transistors for providing linear amplification at 30 Mc. The peak envelope power output is 10 watts. All odd order distortion products on a two tone test are greater than 30 db down from the reference level as measured on a Panoramic SB-12bs Spectrum Analyzer. The overall power gain is approximately 35 db. This performance is easily obtained without using feedback loops. Current feed-back is used in the first two stages for stability by placing unbypassed resistors in the emitters.

Often there is a unique driving source impedance for each transistor which results in the most linear operation. This driving source impedance is obtained in this circuit by the use of torroid transformers having as high a coefficient of coupling as possible and low leakage reactance. The 200 ohm base driving impedance for the first two stages is obtained by using the following simple relationships. T1 provides a 4:1 step-up of the 50-ohm exciter output. T2 is designed to step-up the 20-ohm input resistance of the 2N2950 to 180 ohms required for the collector loading of the 2N2951. Therefore, a 3:1 turns ratio is used. The 2 K resistor across the primary of T-2 sets the maximum base driving impedance of the 2N2950 to 220 ohms assuming no loading due to the collector of the first stage.

The values of biasing shown will set the idle current for each stage at a point which gives the best compromise between peak power output and distortion.

SECTION 12-4

76 MC CW Transmitter

2-WATT CW

A four stage CW transmitter operating in the 72-76 Mc band is shown in Figure 12-4-1. This transmitter provides a 2-watt carrier (rms) from a 25-volt dc supply. The output stage draws 145 mA for a collector efficiency of 55 per cent. The oscillator and driver circuits have a total current drain of 21 mA.

10 WATT CW

This transmitter (see Figure 12-4-2) provides a maximum carrier power of 10 watts from a 25-volt dc supply. The output stage draws 600 mA for a collector efficiency of 66 per cent. The driver stages have a total current drain of 166 mA. The input power required is 6 mW.

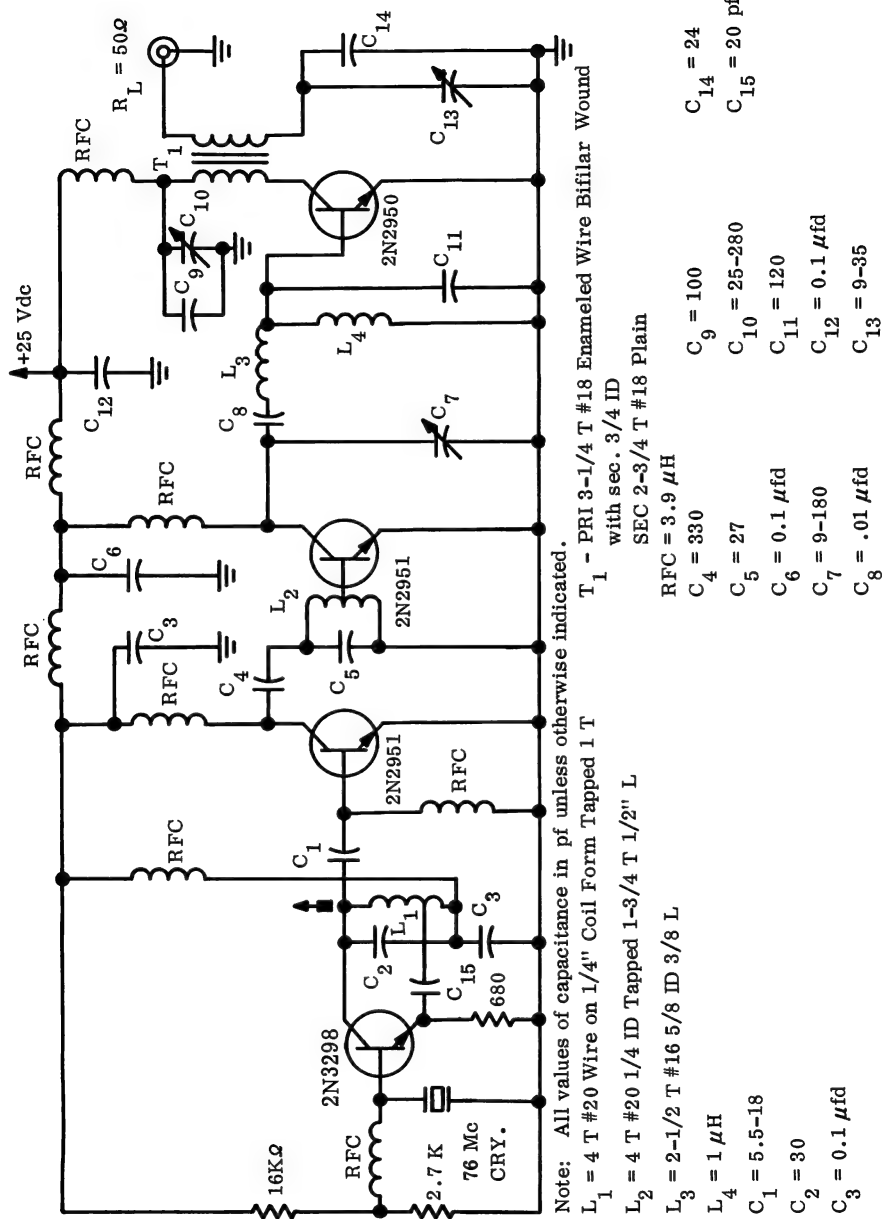
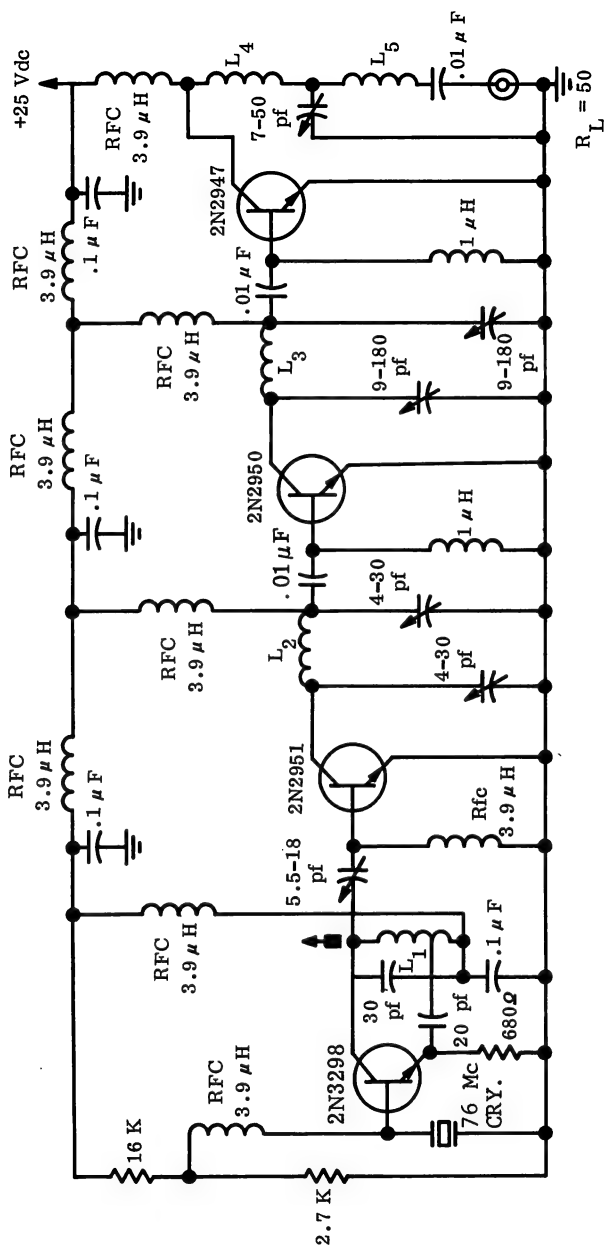


Figure 12-4-1 76 Mc 2-Watt Output



$L_1 = 4 \text{ T \#20, } 1/4" \text{ I.D., Tapped } 1 \text{ T from Ground}$

$L_2, L_3 = 2 \text{ } 1/2 \text{ T \#20, } 1/2" \text{ I.D., } .05 \text{ H}$

$L_4, L_5 = 4 \text{ T \#20, } 1" \text{ I.D., } .15 \text{ H}$

Figure 12-4-2 76 Mc 10-Watt Transmitter

SECTION 12-5

120 MC Aircraft Transmitter

This transmitter (see Figure 12-5-1) will supply 1.5 watts of 120 Mc carrier power (rms) into a 50-ohm load. The total current drain from a 12-volt dc source is approximately 370 mA. The maximum modulation with less than 5 per cent envelope distortion is approximately 75 per cent.

OSCILLATOR

A 60 Mc crystal controlled oscillator is used to generate the 60 Mc carrier. The crystal is operated in the series mode, with feedback provided to the emitter. Driving power is supplied to Q2 by tapping the oscillator coil L1.

PRE-DRIVEN STAGE

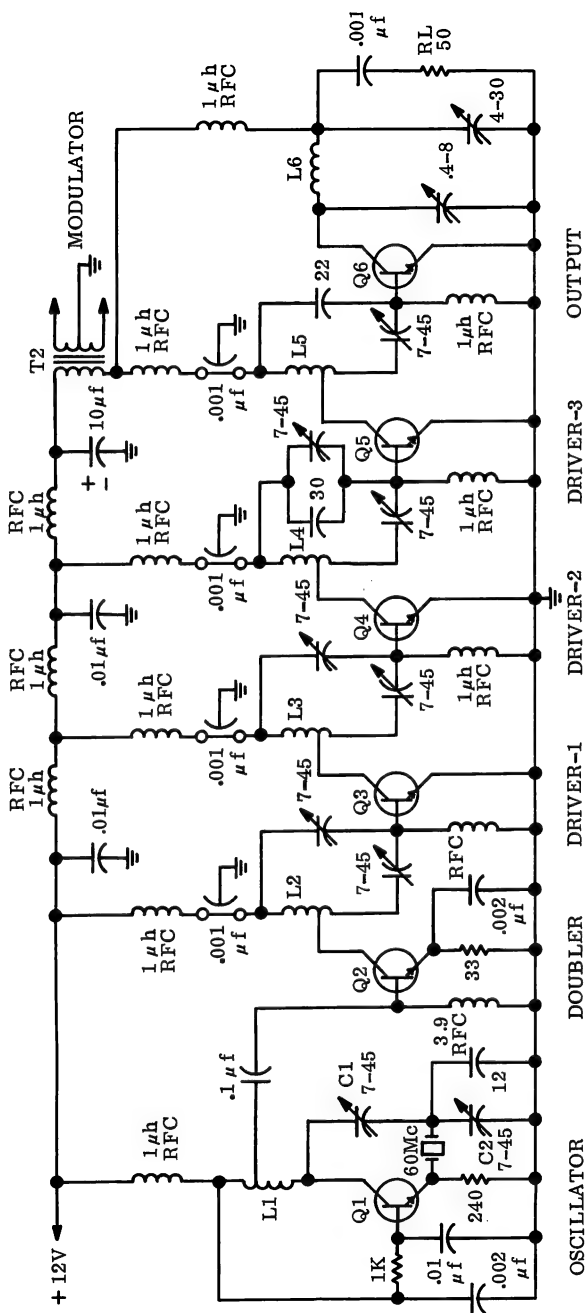
Three stages of amplification using the 2N2951 are necessary to increase the power level to that required at the base of the driver. The second stage (Q2) functions as a doubler circuit also. Detailed coil information is shown in Figure 12-5-1.

DRIVER STAGE

The driver stage is also collector modulated. Modulation of the driver stage is necessary in order to attain 100 per cent modulation (during downward modulation) in the output stage. The maximum downward modulation without driver modulation is approximately 80 per cent.

OUTPUT STAGE

The output stage is collector modulated from a 12-volt dc source. Collector modulation provides the lowest envelope distortion for a 100 per cent modulated carrier, in comparison to base-emitter modulation. The antenna load is simulated by a 50-ohm resistor. This substitution allows the output power to be accurately monitored. Initially, the 50-ohm load is transformed to 48 ohms at the collector by a low-pass pi network. The 48-ohm collector loading will provide 1.5 watts of unmodulated carrier power (rms). However, since a low envelope distortion is desirable, the collector loading is usually adjusted for minimum envelope distortion and maximum device efficiency.



COIL	COIL TURNS	WIRE GAGE	COIL I.D.	IND. μh
L1	6	18	1/4"	.14
L2	1 1/2	14	3/4"	.11
L3	1 1/2	14	3/4"	.11
L4	1 1/2	14	3/4"	.11
L5	1 1/2	14	3/4"	.11
L6	1 1/2	14	3/4"	.11

UNLESS OTHERWISE NOTED:
ALL RESISTORS ARE IN OHMS
ALL CAPACITORS ARE IN MICRO-MICROFARADS

Q1, Q2, Q3, & Q4 - 2N2951
Q5 & Q6 - 2N2950

T2 - THORADSON
TRANSFORMER TR-58

L1 TAPPED 5 TURNS FROM GROUND.
L2, L3, L4, & L5 COLLECTOR
TAPPED 3/4 TURN FROM GROUND.

Figure 12-5-1 120 MC AM Transmitter

SECTION 12-6

A 2-Watt, 160 MC Driver Final

TRANSMITTER

Figure 12-6 is a three stage 160 Mc output circuit designed to deliver 2 watts with an overall gain of 23 db. Q1 is operating Class A and has 10 db gain with an output power of 100 mw. No matching is used on the input since the input impedance of the 2N1142 is about 50 ohms. The small amount of capacitive susceptance on the device input may be parallel tuned out for slightly better performance but this is generally found to not be worth the expenditure of another coil. The collector load impedance is 1000 ohms, which provides fair stability while operating Q1 without neutralization. If a greater stability margin is needed, a coil neutralization scheme will accomplish this. Connect an inductance sufficient to tune approximately 2 pF at 160 Mc from collector-to-base in series with a blocking capacitor. If adjustment is required, use a slightly smaller coil and trim it with a parallel-variable capacitor. A 470-ohm emitter resistor is used to keep Q1 within breakdown voltage ratings.

The MM1306 (Q2) is a germanium high frequency mesa transistor specially selected for 160 Mc driver applications. The 180-ohm emitter resistor prevents the dc voltage across the transistor from exceeding 15 volts. The input and output networks are standard pi-sections and are adjusted for maximum drive to Q3. Zero bias is accomplished with the RF choke from base-to-emitter. This choke should be as small electrically as feasible to minimize the possibility of circuit oscillations at some lower frequency. Typical power output from Q2 is 550 mw.

The 2N2950 (Q3) is an NPN silicon, medium power, RF amplifier transistor which can easily produce 2 watts of RF power at 160 Mc. The typical power gain is about 6 db, with the worst case units having about 5.5 db power gain, depending somewhat upon physical layout and the length of the emitter lead to ground.

SECTION 12-7

240 MC, .5-Watt CW Transmitter

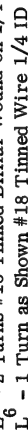
Figure 12-7 is a 4-stage CW transmitter designed to provide a 1/2 watt carrier at 240 Mc from a 12-volt source. It utilizes three silicon NPN annular passivated transistors and one germanium PNP mesa VHF amplifier transistor.

The oscillator is a common-base crystal controlled Hartley type with feedback provided by the 12 pf capacitor. This capacitor allows a practical tapping point on L_1 . Feedback is fixed by the 12 pf capacitor from the emitter to the tap on L_1 , therefore, the only necessary oscillator adjustment is the core of L_1 . The oscillator delivers approximately 15 mw at 80 Mc to the MM1941 tripler.

The bias of the tripler is set by the 75-ohm emitter resistor. The tripler stage provides 30 mw at 240 Mc with 3 db power gain. The performance of this stage is closely related to the bandpass characteristics of the collector network. This network is a double-tuned critically-coupled output circuit tuned to 240 Mc, and provides the necessary selectivity required to reject subharmonics and harmonic spurious outputs of the tripler.

The 30 mw output of the tripler is fed directly to the base of the class "C" MM1943 driver stage. A one-turn tap is used on L_3 to obtain the low driving source impedance required for good class "C" operation of the driver. A simple pi-network couples the output of the driver to the base of the final. The power output of the driver is approximately 150 mw. The dc feed path for the driver is completed through the "zero biasing" RF choke of the final.

The final is a zero biased "class C" stage which is loaded by a double-tuned close-coupled transformer. The primary provides circuit tuning while the secondary capacitance allows the circuit to load to a variety of impedances, roughly from 25 to 100 ohms. The power out data is taken with a 50-ohm load. The minimum output power at 25°C ambient is 500 mw. A typical value is 650 mw. The power output remains above 500 mw in ambient temperatures up to +55°C. Total transmitter drain from a 12-volt dc source is 180 ma at 55°C ambient with the minimum of 500 mw output.



SECTION 12-8

240 MC, 8-Watt CW Transmitter

The transmitter schematic shown in Figure 12-8 utilizes an 80-240 megacycle varactor tripler circuit for the output stage to provide 8 watts of carrier power (rms) at 240 megacycles. The total current drain from a 25-volt dc source is 800 milliamperes.

The drive power at 80 megacycles is provided by a transmitter similar to that shown in Figure 12-4-2. The only circuit change is the replacement of the "T" type network in the output by the tapped double tuned network which is used to drive the varactor.

The varactor tripler circuit consists basically of three tuned circuits: an input circuit, an output circuit and an idler circuit. The output circuit is a double-tuned circuit using capacitive coupling and is tuned to 240 Mc. The idler circuit is a series tuned circuit tuned to 160 Mc. The input circuit, also a double-tuned circuit with capacitive coupling is tuned to 80 Mc. The primary coil is tapped to provide the proper collector loading.

The initial alignment of the varactor circuitry should ideally be made from an 80 Mc, 50-ohm power source. Bandpass filters should be provided on both the input and output circuits to facilitate tuning. If filters are not used, the output waveform should be monitored on a sampling oscilloscope. Initially, the 80 Mc drive source should be adjusted for several watts output. Each tuned circuit may be pre-tuned to the proper frequency with a grid-dip meter. Both coupling capacitors (1-12 pF), should be adjusted to approximately 2 pF. All circuits should be readjusted for maximum output power under drive. When rated power output is realized, the circuit should be driven directly by the transistorized 80 Mc drive source. Retuning of all circuits will be necessary for rated output power and optimum overall efficiency.

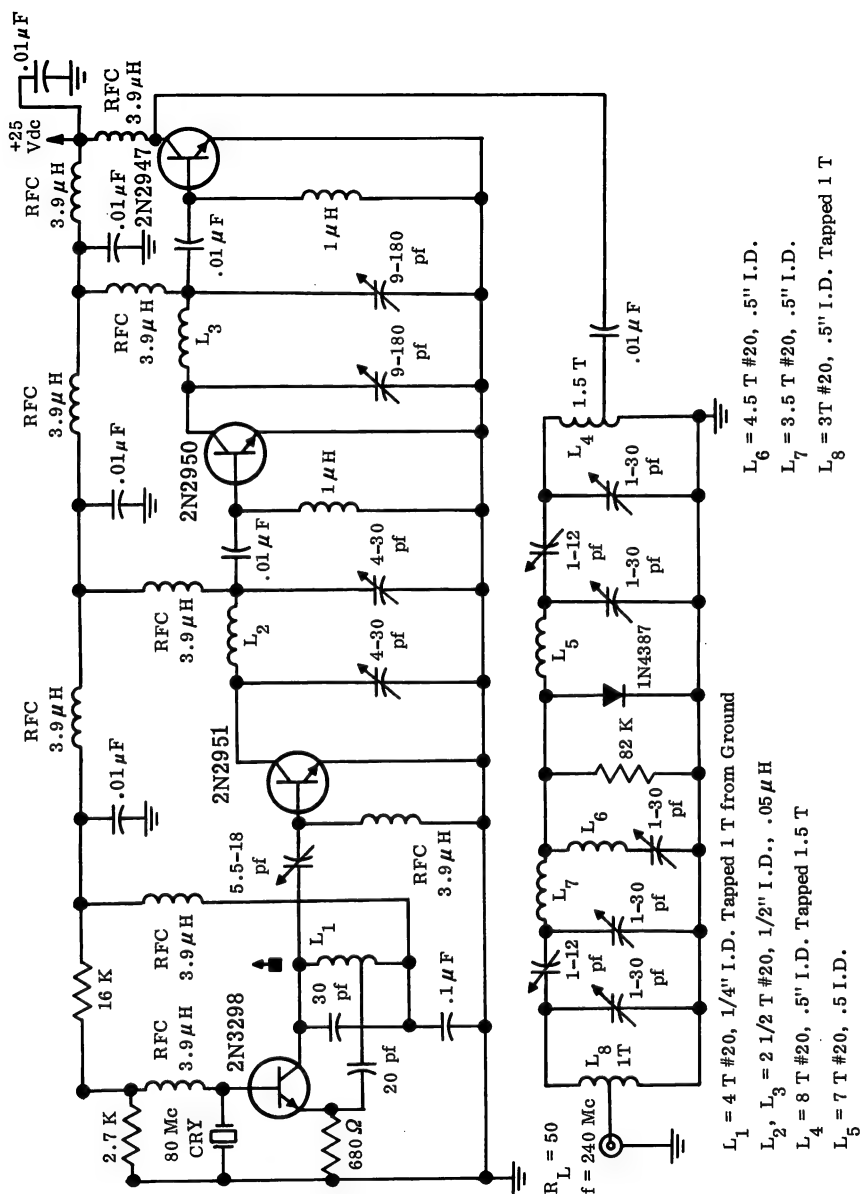


Figure 12-8 240 Mc CW Transmitter with Varactor Tripler

SECTION 13

Device Specifications

The following pages give pertinent design data for the major semiconductor devices referred to in this manual. Complete data sheets for any Motorola Semiconductor Device can be obtained by writing to:

**TECHNICAL INFORMATION CENTER,
MOTOROLA SEMICONDUCTOR PRODUCTS INC.,
P.O. BOX 955 PHOENIX, ARIZONA 85001**

MOTOROLA SILICON CONTROLLED RECTIFIERS

8 AMPERES RMS 25 THRU 400 VOLTS

MCR1304 SERIES



MCR1604 SERIES



MCR1305 SERIES



MCR1605 SERIES



These silicon controlled rectifiers are particularly applicable for power control and switching circuits such as; motor speed control, lamp dimming, flashers, battery charger regulators, power inverters and similar circuits requiring currents up to 8-amperes.

MAXIMUM RATINGS ($T_J = 100^{\circ}\text{C}$ unless otherwise noted)

Characteristic	Symbol	Rating	Unit
Peak Forward Voltage	PFV	600	Volts
Peak Reverse Blocking Voltage* <div style="display: flex; align-items: center;"> <div style="margin-right: 10px;"> MCR1304 -1 MCR1305 -2 MCR1604 -3 MCR1605 -4 MCR1605 -5 MCR1605 -6 </div> <div style="font-size: 3em; margin-right: 10px;">}</div> </div>	$V_{ROM(rep)}$	25 50 100 200 300 400	Volts
Forward Current RMS (All Conduction Angles)	I_F	8	Amps
Peak Surge Forward Current (One cycle, 60 cps) ($T_J = -40$ to $+100^{\circ}\text{C}$)	$I_{FM(surge)}$	100	Amps
Circuit Fusing Considerations ($T_J = -40$ to $+100^{\circ}\text{C}$; ≤ 8.3 msec)	I^2t	40	A^2sec
Peak Gate Power	P_{GM}	5	Watts
Average Gate Power	$P_{G(av)}$	0.5	Watts
Peak Gate Current	I_{GFM}	2	Amps
Peak Gate Voltage Forward Reverse	V_{GFM} V_{GRM}	10 10	Volts
Storage Temperature Range	T_{stg}	-40 to $+150$	$^{\circ}\text{C}$
Operating Temperature Range	T_J	-40 to $+100$	$^{\circ}\text{C}$
Stud Torque (MCR1305 series)		15	in. lb.

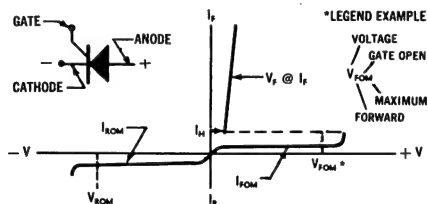
* $V_{ROM(rep)}$ for all types can be applied on a continuous DC basis without incurring damage. Ratings apply for zero or negative gate voltage.

ELECTRICAL CHARACTERISTICS (T_J = 25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Voltage* (T _J = 100°C)	V _{FOM} *	25 50 100 200 300 400	— — — — — —	— — — — — —	Volts
Peak Forward Blocking Current (Rated V _{FOM} @ T _J = 100°C, gate open)	I _{FOM}	—	—	2	mA
Peak Reverse Blocking Current (Rated V _{ROM} @ T _J = 100°C)	I _{ROM}	—	—	2	mA
Forward On Voltage (I _F = 5 Adc)	V _F	—	1.0	1.3	Volts
Gate Triggering Current (Anode Voltage = 7 Vdc, R _L = 100 Ω)	I _{GT}	—	7	20	mA
Gate Triggering Voltage (Anode Voltage = 7 Vdc, R _L = 100 Ω) (Anode Voltage = 7 Vdc, R _L = 100 Ω, T _J = 100°C)	V _{GT} V _{GNT}	— 0.2	0.6 —	1.5 —	Volts
Holding Current (Anode Voltage = 7 Vdc, gate open)	I _{HO}	—	8	25	mA
Turn-On Time (I _F = 5 Adc, I _{GT} = 20 mAdc)	t _{on}	—	1	—	μsec
Turn-Off Time (I _F = 5 Adc, I _R = 5 Adc) (I _F = 5 Adc, I _R = 5 Adc, T _J = 100°C)	t _{off}	— —	12 16	— —	μsec
Forward Voltage Application Rate (T _J = 100°C)	dv/dt	—	50	—	V/μsec
Thermal Resistance	θ _{JC}	— —	1.5 —	2.7 3.0	°C/W
Thermal Resistance	θ _{CA}	—	—	40	°C/W

* V_{FOM} for all types can be applied on a continuous DC basis without incurring damage.
Ratings apply for zero or negative gate voltage.

SCR SYMBOLS and DEFINITIONS



TYPICAL E-I CHARACTERISTICS

dv/dt

FORWARD VOLTAGE APPLICATION RATE (V/μsec) - A rate of applied voltage in excess of this value may cause premature, non-destructive breakover.

I_F

FORWARD CURRENT - The RMS value of forward current during the "on" state. RMS value is the same for all conduction angles.

I_{F(av)}

AVERAGE FORWARD CURRENT - The full cycle average forward current for specified conditions of case temperature and conduction angle.

I_{FM}

PEAK FORWARD CURRENT, "ON" STATE - The peak current through the collector junction for a positive anode-to-cathode voltage.

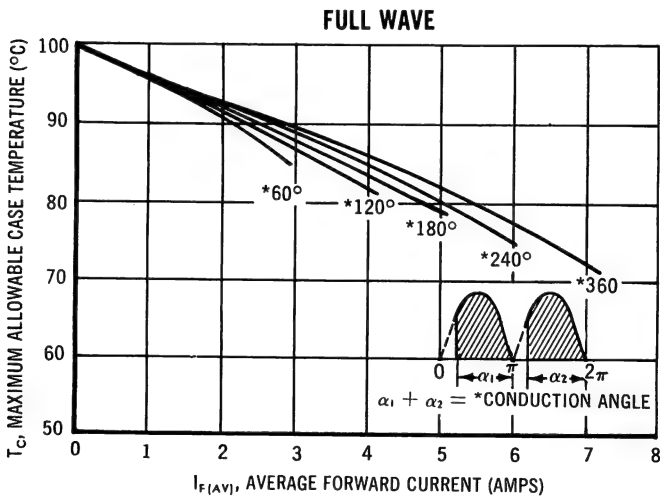
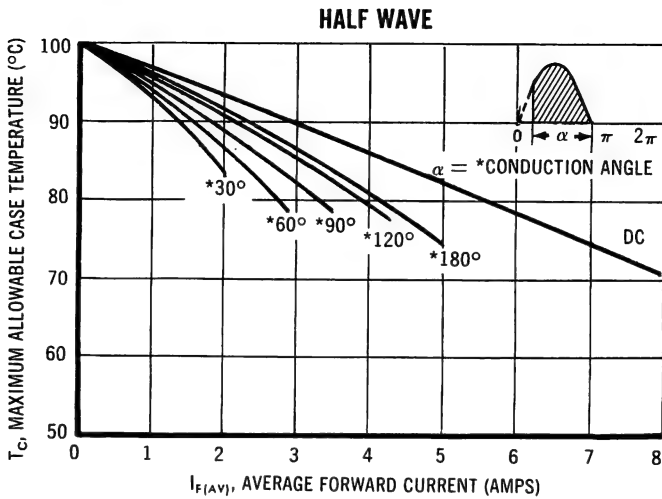
I_{FM(surge)}

PEAK ONE-CYCLE SURGE FORWARD CURRENT - The maximum forward current having a single forward cycle (8.3 milliseconds duration) in a 60 cps single-phase resistive load system. The surge may be preceded and followed by maximum rated voltage, current, and junction temperature conditions, and maximum allowable gate power may be concurrently dissipated. However, limitations on anode current during turn-on should not be exceeded.

I_{FOM}	PEAK FORWARD BLOCKING CURRENT, GATE OPEN - The peak current through the collector junction when the SCR is in the "off" state for a stated anode - to - cathode voltage and junction temperature.
I_{GT}	GATE TRIGGER CURRENT, DC - The minimum DC gate current required to cause switching from the "off" state to the "on" state for a stated anode-to-cathode voltage.
I_{HO}	HOLDING CURRENT, GATE OPEN - That value of forward anode current below which the controlled rectifier switches from the conducting state to the forward blocking condition.
I_{ROM}	PEAK REVERSE BLOCKING CURRENT, GATE OPEN - The peak current through the collector junction when the SCR is in the reverse blocking state for a stated anode - to - cathode voltage and junction temperature.
I_t^2	(FOR CIRCUIT FUSING CONSIDERATIONS) - A measure of the maximum non-recurrent RMS forward current capacity for pulse durations less than 8.3 milliseconds. I is in RMS amperes, and t is pulse duration in seconds. The same conditions as listed for $I_{FM(surge)}$ apply.
PFV	PEAK FORWARD VOLTAGE - The maximum non-destructive value of forward voltage which may be applied to any device in the series, with the gate open. The PFV value should not be exceeded even though a unit may intrinsically, or under certain conditions, be capable of blocking much higher voltages.
$P_{G(av)}$	AVERAGE GATE POWER DISSIPATION - The value of maximum gate power dissipation averaged over a full cycle permitted between gate and cathode.
P_{GM}	PEAK GATE POWER DISSIPATION - The maximum instantaneous value of gate power dissipation permitted between gate and cathode.
R_L	LOAD RESISTOR
T_C	CASE TEMPERATURE
T_J	JUNCTION TEMPERATURE
T_{stg}	STORAGE TEMPERATURE
t_{on}	TURN-ON TIME - The time interval between initiation of the gate current signal and reduction of the forward voltage to 10% of the blocking value during switching to conduction under stated conditions.
t_{off}	TURN-OFF TIME - The time interval required for the gate to regain control of forward blocking characteristic after interruption of forward anode current.
V_{FM}	PEAK "ON" VOLTAGE - The peak forward voltage for a stated peak forward current when the SCR is in the "on" state.
V_{FOM}	PEAK FORWARD BLOCKING VOLTAGE, GATE OPEN - The peak forward voltage when the SCR is in the "off" state.
V_{GFM}	PEAK FORWARD GATE VOLTAGE - The peak voltage between the gate terminal and the cathode terminal resulting from the flow of forward gate current.
V_{GRM}	PEAK REVERSE GATE VOLTAGE - The peak voltage between the gate terminal and the cathode terminal when the junction between the gate region and the adjacent cathode region is reverse biased.
V_{GT}	GATE TRIGGER VOLTAGE, DC - The DC voltage between the gate and the cathode required to produce the DC gate trigger current.
$V_{ROM(rep)}$	PEAK REVERSE VOLTAGE, GATE OPEN - The maximum allowable instantaneous value of reverse voltage (repetitive or continuous DC) which can be applied to the device with the gate open at rated temperature.
θ_{JC}	THERMAL RESISTANCE, JUNCTION-TO-CASE - The temperature rise per unit power dissipation of a designated junction above the temperature of the case under conditions of thermal equilibrium.

CURRENT DERATING *

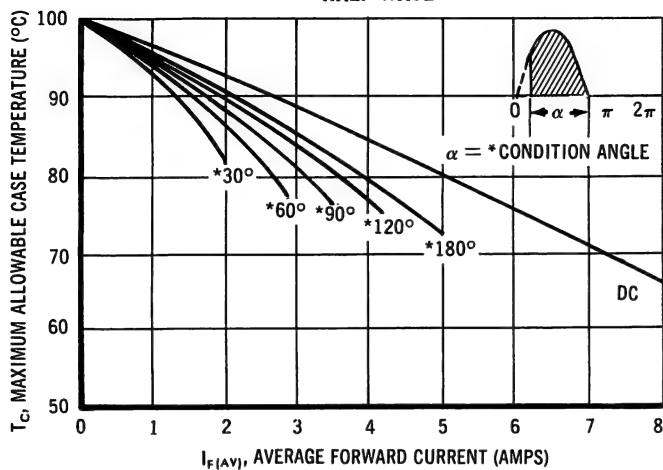
MCR1304, MCR1604, and MCR1605 series



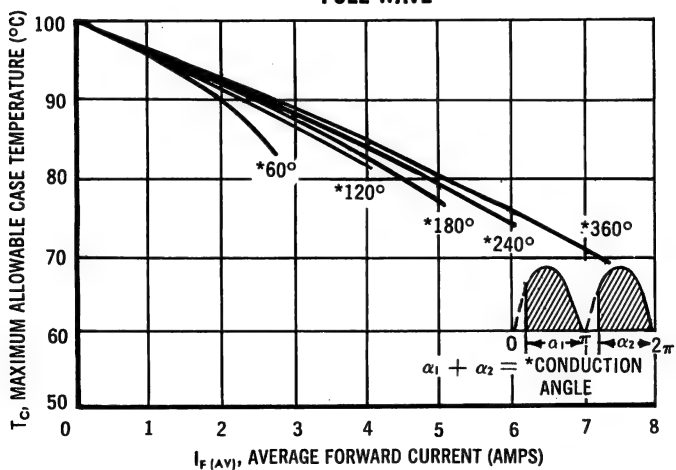
*Power dissipation curves are based upon the maximum forward voltage versus forward current at 100° C junction temperature. Current derating curves are based upon maximum power dissipation curves and the coefficient of power of the SCR. The coefficient of power takes into consideration the thermal response of the SCR to power pulses as explained in section 1-5 of the Motorola Zener Diode and Rectifier Handbook.

MCR1305 series

HALF WAVE

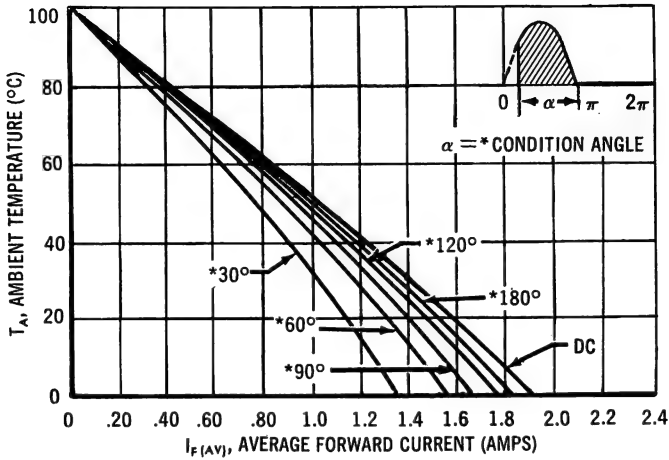


FULL WAVE

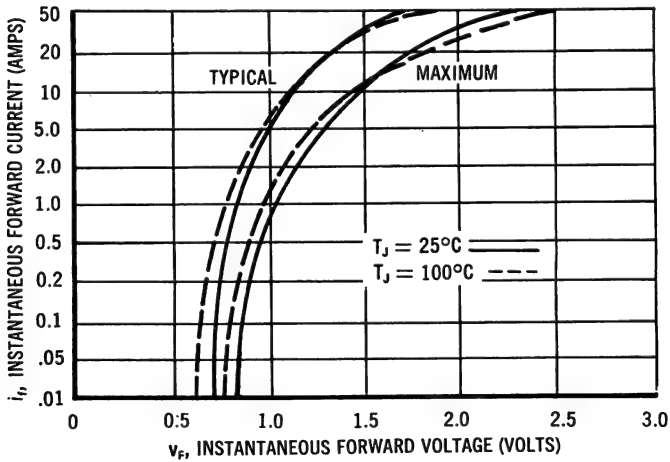


MCR1304, MCR1604, and MCR1605

Free air mounting, anode and cathode leads terminated at end points.

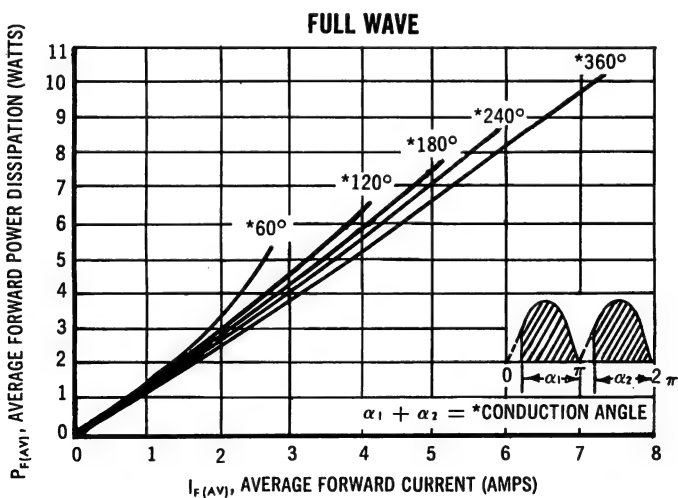
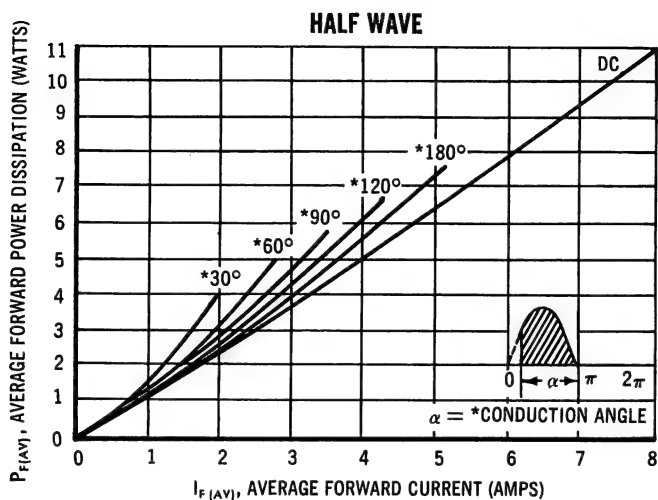


FORWARD CONDUCTING CHARACTERISTICS ALL TYPES

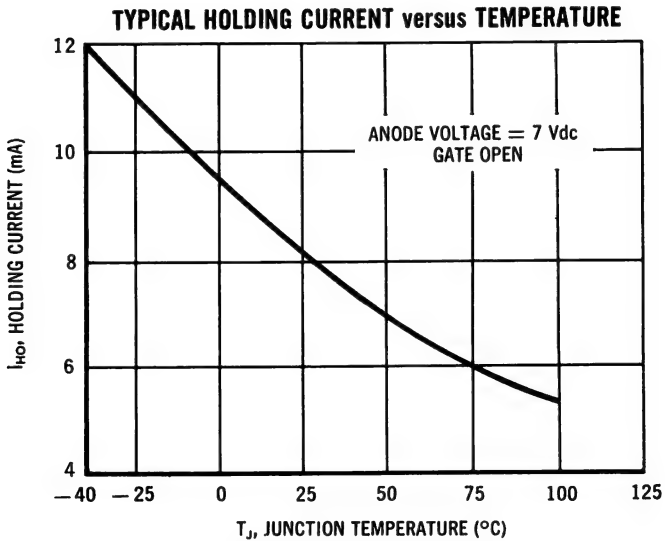
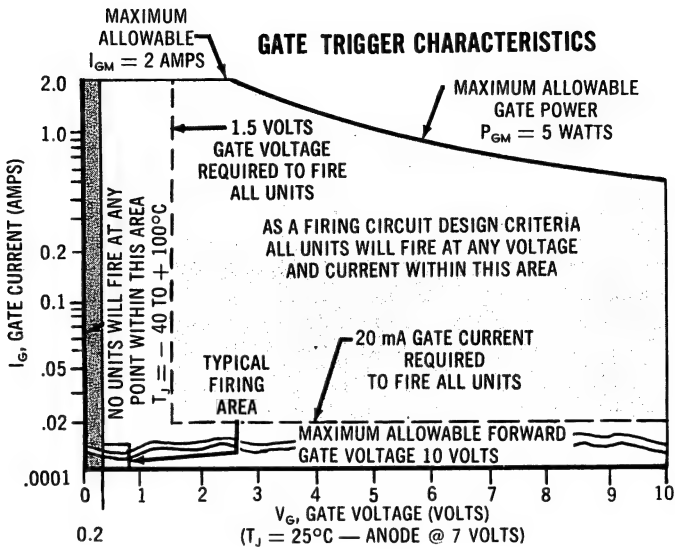


FORWARD POWER DISSIPATION

ALL TYPES



GATE CHARACTERISTICS

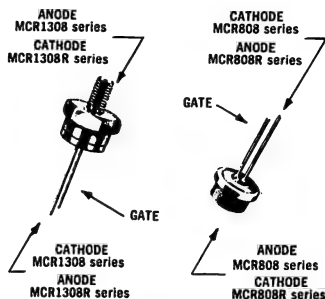


MOTOROLA SILICON CONTROLLED RECTIFIERS

18 AMPERES RMS 25 THRU 400 VOLTS

**MCR808 SERIES
MCR1308 SERIES
(FORWARD POLARITY)**

**MCR808R SERIES
MCR1308R SERIES
(REVERSE POLARITY)**



The MCR808 Series of silicon controlled rectifiers is capable of handling RMS currents up to 18 amperes. These devices are ideal for motor speed controllers, lamp dimmers and flashers and other power-control applications.

MAXIMUM RATINGS ($T_J = 100^\circ\text{C}$ unless otherwise noted)

Characteristic	Symbol	Rating	Unit
Peak Forward Voltage	PFV	600	Volts
Peak Reverse Voltage* <div style="display: flex; align-items: center; justify-content: center;"> <div style="margin-right: 10px;"> MCR808 MCR808R MCR1308 MCR1308R </div> <div style="font-size: 3em; margin-right: 10px;">}</div> <div style="display: flex; flex-direction: column; align-items: center;"> -1 -2 -3 -4 -5 -6 </div> </div>	$V_{ROM(rep)}$	25 50 100 200 300 400	Volts
Forward Current RMS (All Conduction Angles)	I_f	18	Amps
Peak Surge Current (one cycle, 60 cps) ($T_J = -40$ to $+100^\circ\text{C}$)	$I_{FM(surge)}$	225	Amps
Circuit Fusing Considerations ($T_J = -40$ to $+100^\circ\text{C}$; 8.3 msec)	I^2t	235	A^2sec
Peak Gate Power	P_{GM}	5	Watts
Average Gate Power	$P_{G(av)}$	0.5	Watts
Peak Forward Gate Current	I_{GFM}	2	Amps
Peak Gate Voltage Forward Reverse	V_{GFM} V_{GRM}	10 10	Volts
Operating Temperature Range	T_J	-40 to +100	$^\circ\text{C}$
Storage Temperature Range	T_{stg}	-40 to +150	$^\circ\text{C}$
Stud Torque - MCR1308 and MCR1308R	—	30	in. lb.

* $V_{ROM(rep)}$ for all types can be applied on a continuous DC basis without incurring damage. Ratings apply for zero or negative gate voltage.

ELECTRICAL CHARACTERISTICS (T_J = 25°C unless otherwise noted)

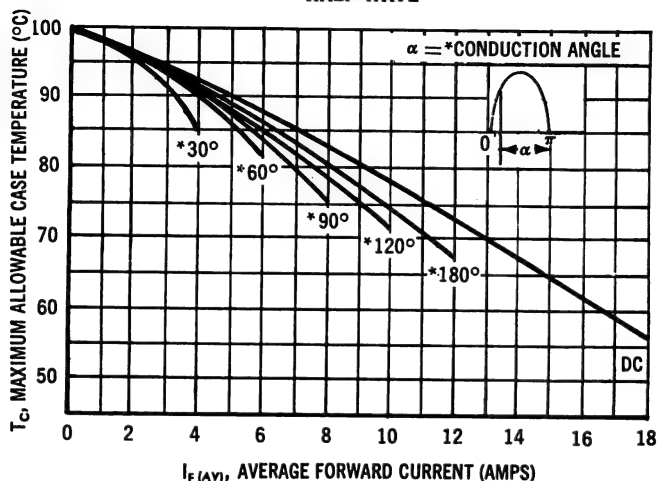
Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Voltage * (T _J = 100°C) <div> MCR808 -1 MCR808R -2 MCR1308 -3 MCR1308R -4 -5 -6 </div>	V _{FOM}	25 50 100 200 300 400	— — — — —	— — — — —	Volts
Peak Forward Blocking Current (Rated V _{FOM} with gate open, T _J = 100°C)	I _{FOM}	—	1	8	mA
Peak Reverse Blocking Current (Rated V _{ROM} , Gate open, T _J = 100°C)	I _{ROM}	—	1	5	mA
Forward On Voltage (18 Adc, T _J = 100°C)	V _F	—	1.1	1.5	Volts
Gate Triggering Current (Anode Voltage = 7 V, R _L = 50 Ω)	I _{GT}	—	10	50	mA
Gate Firing Voltage (Continuous DC) (Anode Voltage = 7 V, R _L = 50 Ω) (Anode Voltage = 7 V, R _L = 50 Ω, T _J = 100°C)	V _{GT} V _{GNT}	— 0.3	0.7 —	1.5 —	Volts
Holding Current (Anode Voltage = 7 V, gate open)	I _{HO}	—	8	—	mA
Turn-On Time (I _F = 18 Adc, I _{GT} = 50 mAdc)	t _{on}	—	1.0	—	μsec
Turn-Off Time (I _F = 10 A, I _R = 10 A) (I _F = 10 A, I _R = 10 A, T _J = 100°C)	t _{off}	— —	12 18	— —	μsec
Forward Voltage Application Rate (T _J = 100°C)	dv/dt	—	30	—	V/μsec
Thermal Resistance MCR808 and MCR808R MCR1308 and MCR1308R	θ _{JC}	—	1.0 1.4	1.6 2.0	°C/W

*V_{FOM} for all types can be applied on a continuous DC basis without incurring damage.
 Ratings apply for zero or negative gate voltage.

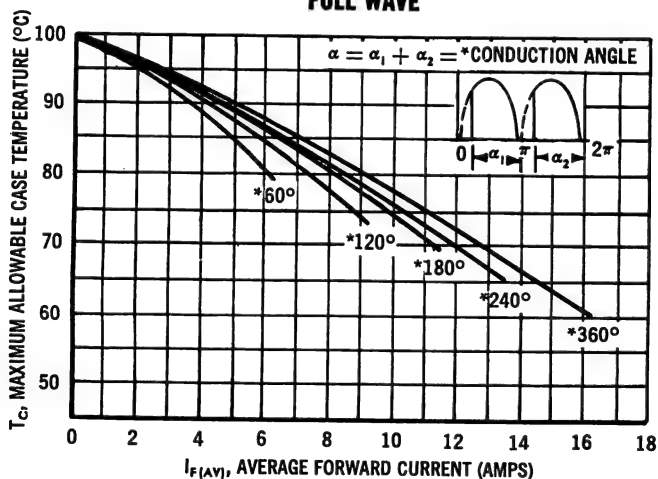
CURRENT DERATING*

MCR808 and MCR808R series

HALF WAVE



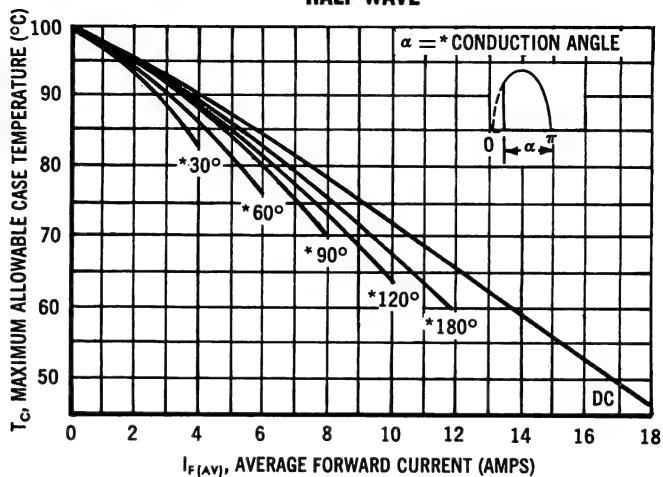
FULL WAVE



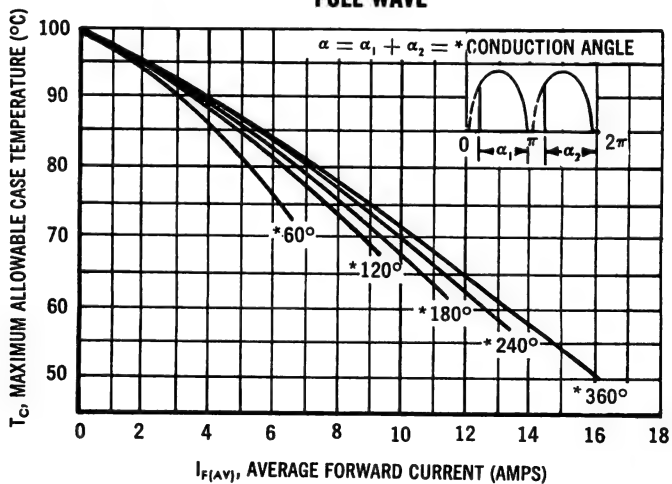
*Power dissipation curves are based upon the maximum forward voltage versus forward current at 100°C junction temperature. Current derating curves are based upon maximum power dissipation curves and the coefficient of power of the SCR. The coefficient of power takes into consideration the thermal response of the SCR to power pulses as explained in section 1-5 of the Motorola Zener Diode and Rectifier Handbook.

MCR1308 and MCR1308R series

HALF WAVE



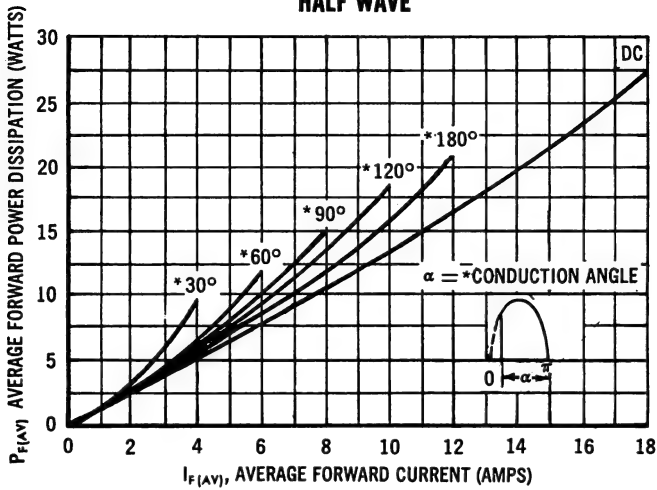
FULL WAVE



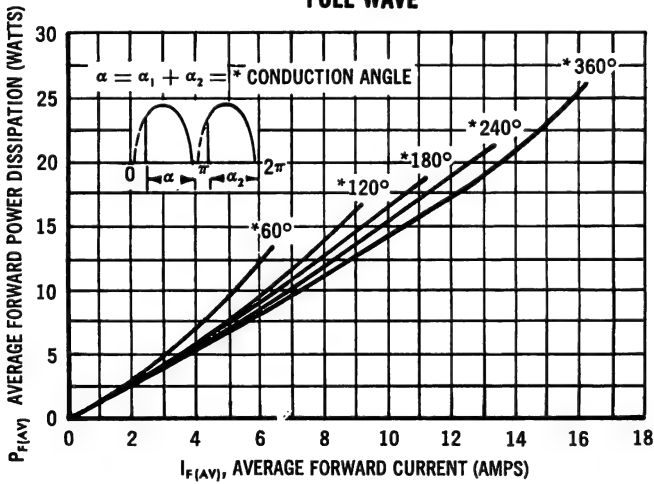
POWER DISSIPATION

MCR808, MCR808R, MCR1308, and MCR1308R series

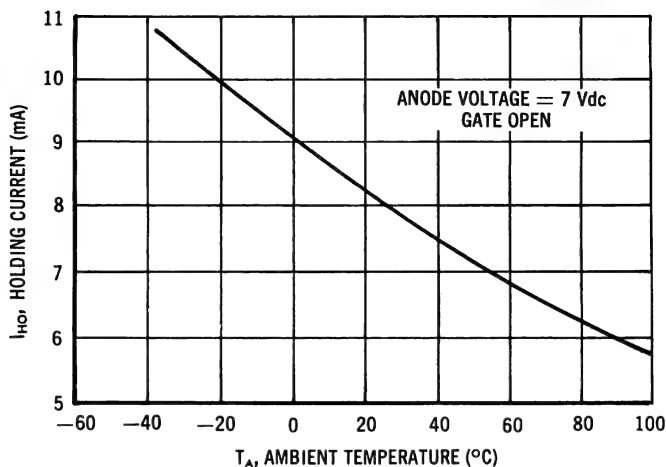
HALF WAVE



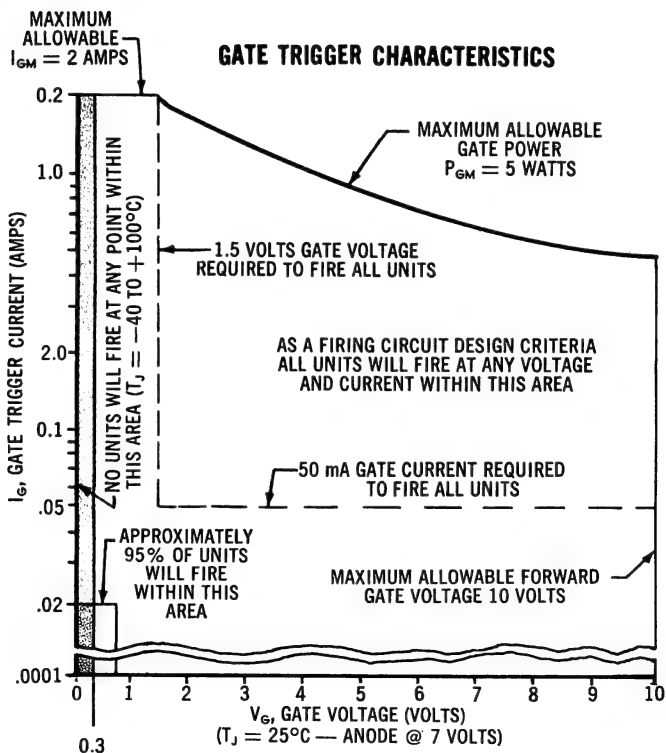
FULL WAVE



TYPICAL HOLDING CURRENT versus TEMPERATURE



GATE TRIGGER CHARACTERISTICS



MOTOROLA 20-AMPERE POWER TRANSISTORS

GERMANIUM PNP EPITAXIAL BASE

2N2832 THRU 2N2834

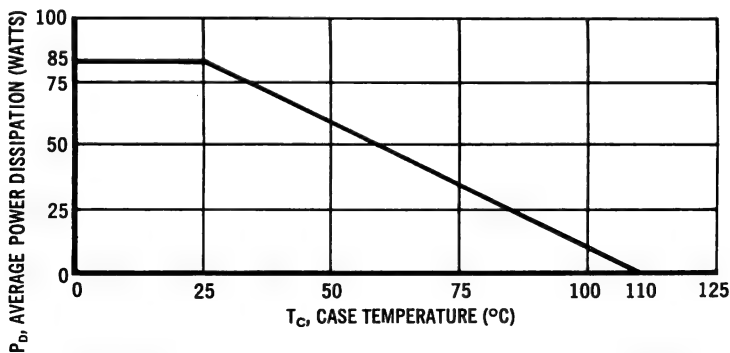


The 2N2832-34 power transistors are high-speed, high-frequency devices especially recommended for applications such as TV horizontal output stages, marine band "class C" power amplifiers and high-frequency power inverters.

ABSOLUTE MAXIMUM RATINGS

Characteristic	Symbol	2N2832	2N2833	2N2834	Unit
Collector-Base Voltage	V_{CBO}	80	120	140	Volts
Collector-Emitter Voltage	V_{CEO}	50	75	100	Volts
Emitter-Base Voltage	V_{EBO}	2	* 2	2	Volts
Collector Current (Continuous)	I_C	20	20	20	Amps
Base Current (Continuous)	I_B	5	5	5	Amps
Power Dissipation	P_C	Fig. 1, 2	Fig. 1, 3	Fig. 1, 4	Watts
Junction Operating Temperature Range	T_j	-65 to +110°			°C

FIGURE 1 — POWER DERATING CURVE



THESE TRANSISTORS ARE ALSO SUBJECT TO SAFE AREA CURVES AS INDICATED BY FIGURES 2, 3, 4. BOTH LIMITS ARE APPLICABLE AND MUST BE OBSERVED

ELECTRICAL CHARACTERISTICS (at $T_C = 25^\circ\text{C}$ unless otherwise noted)

Characteristic	Symbol	Min	Typical	Max	Unit
Collector-Base Cutoff Current* $(V_{CB} = 2V, I_E = 0)$ All Types $(V_{CB} = 80V, I_E = 0)$ 2N2832 $(V_{CB} = 120V, I_E = 0)$ 2N2833 $(V_{CB} = 140V, I_E = 0)$ 2N2834	I_{CBO}^*	---	---	0.3 10 10 10	mA
Collector-Emitter Current* $(V_{CE} = 100V, V_{BE} = 0)$ 2N2832 $(V_{CE} = 140V, V_{BE} = 0)$ 2N2833 $(V_{CE} = 160V, V_{BE} = 0)$ 2N2834	I_{CES}^*	---	---	20 20 20	mA
Collector-Emitter Cutoff Current** $(V_{CE} = 50V, V_{BE} = 0.2V, T_C = +85^\circ\text{C})$ 2N2832 $(V_{CE} = 75V, V_{BE} = 0.2V, T_C = +85^\circ\text{C})$ 2N2833 $(V_{CE} = 100V, V_{BE} = 0.2V, T_C = +85^\circ\text{C})$ 2N2834	I_{CEX}^{**}	---	---	40 40 40	mA
Emitter-Base Breakdown Voltage $(I_E = 50 \text{ mAdc}, I_C = 0)$	BV_{EBO}	2	---	---	Vdc
Collector-Emitter Breakdown Voltage** $(I_E = 100 \text{ mA}, I_B = 0)$ 2N2832 2N2833 2N2834	$BV_{CEO(sus)}^{**}$	50 75 100	---	---	Volts
Emitter Floating Potential* $(V_{CB} = 80V, I_E = 0)$ 2N2832 $(V_{CB} = 120V, I_E = 0)$ 2N2833 $(V_{CB} = 140V, I_E = 0)$ 2N2834	$VEBF^*$	---	---	0.5 0.5 0.5	Volts
DC Current Transfer Ratio $(I_C = 1.0 \text{ A}, V_{CB} = 2V)$ $(I_C = 10 \text{ A}, V_{CB} = 2V)$	h_{FE}	50 25	75 ---	---	---
Collector-Emitter Saturation Voltage $(I_C = 1.0 \text{ Adc}, I_B = 100 \text{ mAdc})$ $(I_C = 10 \text{ Adc}, I_B = 1.0 \text{ Adc})$ $(I_C = 20 \text{ Adc}, I_B = 2.0 \text{ Adc})$	$V_{CE(sat)}$	---	---	0.15 0.30 0.5	Vdc
Base-Emitter Saturation Voltage $(I_C = 1 \text{ A}, I_B = 100 \text{ mAdc})$ $(I_C = 10 \text{ A}, I_B = 1 \text{ Adc})$ $(I_C = 20 \text{ A}, I_B = 2 \text{ Adc})$	$V_{BE(sat)}$	---	---	0.6 0.75 1.0	Vdc
Small Signal Current Gain $(I_C = 1.0 \text{ A}, V_{CE} = 10 \text{ V}, f = 5 \text{ mc})$	h_{fe}	2	3.5	---	---
Rise Time	t_r	---	2	4	μsec
Storage Time	t_s	---	3	6	μsec
Fall Time	t_f	---	1	2.5	μsec

* SWEEP TEST: 1/2 Sine Wave, 60 cps min

** PULSE TEST: PW = 1 msec, 2% Duty Cycle

SAFE OPERATING AREAS

FIGURE 2 — 2N2832

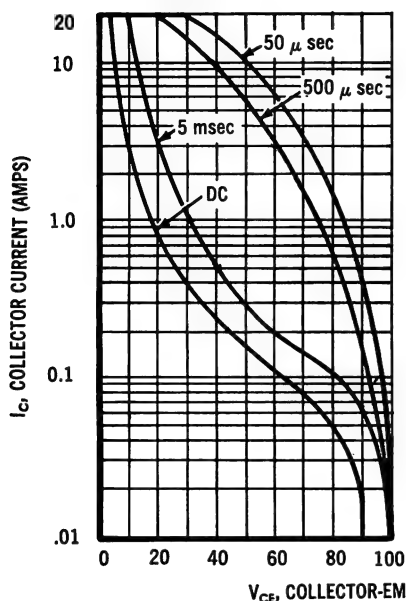


FIGURE 3 — 2N2833

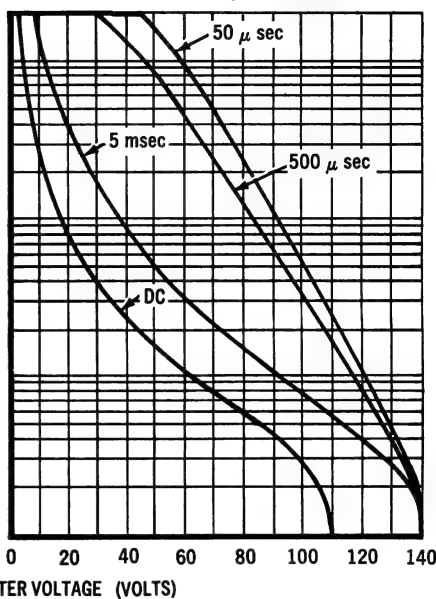
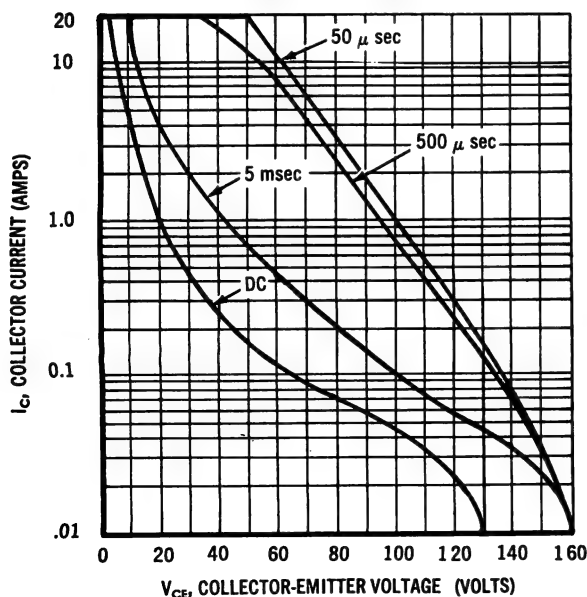


FIGURE 4 — 2N2834



The Safe Operating Area Curves indicate the $I_c - V_{ce}$ limits below which the devices will not go into secondary breakdown. As secondary breakdown is independent of temperature and duty cycle, these curves can be used as long as the average power derating curve (Figure 1) is also taken into consideration to insure operation below the maximum junction temperature.

FIGURE 5 — CURRENT-GAIN VARIATIONS

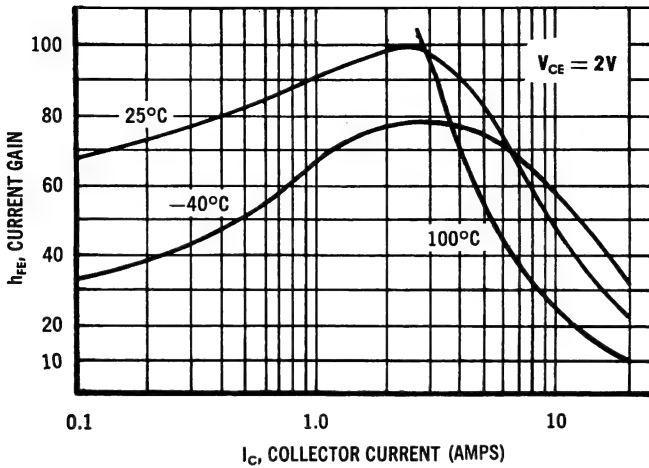
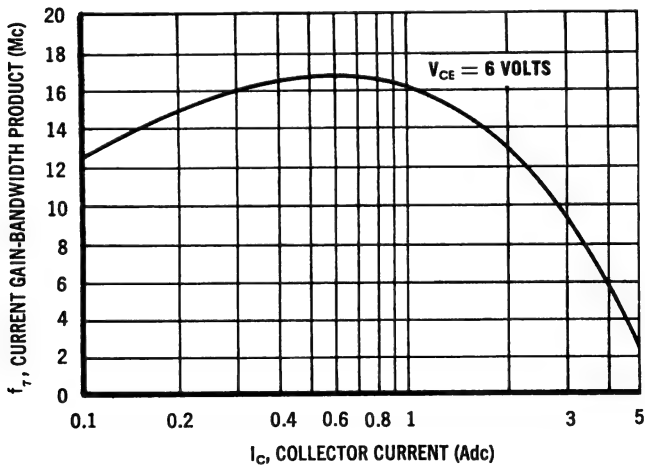


FIGURE 6 — CURRENT GAIN-BANDWIDTH PRODUCT vs COLLECTOR CURRENT



MOTOROLA 25- AMPERE POWER TRANSISTOR

GERMANIUM PNP EPITAXIAL BASE

2N2912



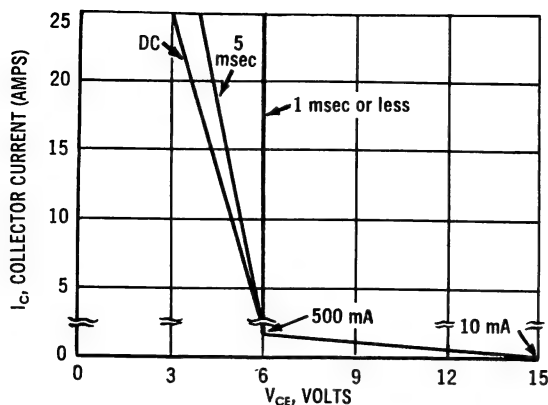
The 2N2912 is particularly designed for high-efficiency power converter circuits operating from minimal voltage sources such as 1.5-V batteries, solar cells, thermo-electric generators and fuel cells.

ABSOLUTE MAXIMUM RATINGS

Characteristic	Symbol	Rating	Unit
Collector-Base Voltage	V_{CBO}	15	Volts
Collector-Emitter Voltage	V_{CEO}	6	Volts
Emitter-Base Voltage	V_{EBO}	1.5	Volts
Collector Current (Continuous)	I_C	25	Amps
Base Current (Continuous)	I_B	3	Amps
Power Dissipation	P_D	See Fig. 1	Watts
Operating Case Temperature Range	T_C	-65 to +110	$^{\circ}\text{C}$
Storage Temperature Range	T_{stg}	-65 to +110	$^{\circ}\text{C}$

THERMAL RESISTANCE: $\theta_{JC} = 1^{\circ}\text{C}/\text{W}$ (junction to case)
 $\theta_{CA} = 30^{\circ}\text{C}/\text{W}$ (case to ambient)

THERMAL TIME CONSTANT: $\tau_{CA} = 1$ minute (case to ambient)



The safe operating area curve is applicable for all case temperatures. The thermal derating curve shown below must also be observed to insure operation below maximum T_J .

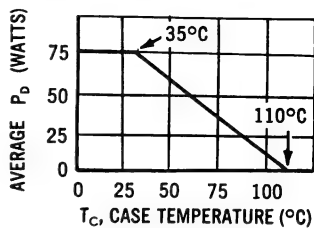


FIGURE 1 — SAFE AREA OF OPERATION

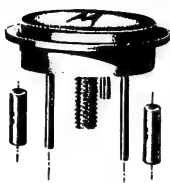
ELECTRICAL CHARACTERISTICS (at $T_C = 25^\circ\text{C}$ unless otherwise noted)

Characteristic	Symbol	Min	Max	Unit
Collector Cutoff Current ($V_{CE} = 15\text{ Vdc}$, $V_{BE} = 0.2\text{ Vdc}$) ($V_{CE} = 6\text{ Vdc}$, $V_{BE} = 0.2\text{ Vdc}$, $T_C = 85^\circ\text{C}$)	I_{CEX}	---	10 15	mAdc
Collector-Base Cutoff Current ($V_{CB} = 2\text{ Vdc}$, $I_E = 0$)	I_{CBO}	---	0.2	mAdc
Emitter Floating Potential ($V_{CB} = 15\text{ Vdc}$, $I_E = 0$)	V_{EBF}	---	0.2	Vdc
Emitter Cutoff Current ($V_{EB} = 1.5\text{ Vdc}$)	I_{EBO}	---	50	mAdc
Collector-Emitter Voltage* ($I_C = 500\text{ mAdc}$, $I_B = 0$)	V_{CEO}^*	6	---	Vdc
DC Forward Current Transfer Ratio ($V_{CE} = 2\text{ Vdc}$, $I_C = 10\text{ Adc}$)	h_{FE}	75	---	---
Collector-Emitter Saturation Voltage ($I_C = 5\text{ Adc}$, $I_B = 0.5\text{ Adc}$) ($I_C = 25\text{ Adc}$, $I_B = 2.5\text{ Adc}$)	$V_{CE}(\text{sat})$	---	0.12 0.5	Vdc
Base-Emitter Saturation Voltage ($I_C = 25\text{ Adc}$, $I_B = 2.5\text{ Adc}$)	$V_{BE}(\text{sat})$	---	1.2	Vdc
Base-Emitter Saturation Voltage ($I_C = 5\text{ Adc}$, $I_B = 0.5\text{ Adc}$)	$V_{BE}(\text{sat})$	---	0.5	Vdc
Rise Time ($I_C = 5\text{ A}$, $V_{CE} = 10\text{ V}$, $I_{B(\text{on})} = 250\text{ mA}$)	t_r	---	2	μsec
Storage Time ($I_C = 5\text{ A}$, $V_{CE} = 10\text{ V}$, $I_{B(\text{on})} = 250\text{ mA}$)	t_s	---	2.5	μsec
Fall Time ($I_C = 5\text{ A}$, $V_{CE} = 10\text{ V}$, $V_{EB} = 1.5\text{ V}$, $R_{be} = 6\text{ Ohms}$)	t_f	---	2	μsec
Common Emitter Small-Signal Forward Current Transfer Ratio Cutoff Frequency ($V_{CE} = 2\text{ V}$, $I_C = 5\text{ A}$, $f = 1\text{ mc}$)	$ h_{fe} $	20	---	---

*Sweep Test

MOTOROLA POWER TRANSISTOR
50- AMPERE, 15- VOLT, 170-WATT
GERMANIUM ALLOY PNP

2N2728



High-current power transistor especially designed for switching and power converter circuits operating from low-voltage power source such as solar cells, thermo-electric generators, sea cells, fuel cells and 1.5-volt batteries.

ABSOLUTE MAXIMUM RATINGS

Characteristic	Symbol	Rating	Unit
Collector-Base Voltage	BV_{CBO}	15	Vdc
Collector-Emitter Voltage	BV_{CEO}	5	Vdc
Emitter-Base Voltage	BV_{EBO}	15	Vdc
Collector Current (continuous)	I_C	50	Adc
Base Current (continuous)	I_B	10	Adc
Total Device Dissipation @ 25°C Case Temperature	P_D	170	Watts
Operating Temperature	T_J	+110	°C
Storage Temperature	T_{stg}	-65 to +110	°C
Thermal Resistance (Junction to Case)	θ_{JC}	0.5	°C/W

ELECTRICAL CHARACTERISTICS (At 25°C unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Collector Cutoff Current $V_{CE} = 15 \text{ V}, V_{BE} = 1 \text{ V}$ $V_{CE} = 10 \text{ V}, V_{BE} = 1 \text{ V}, T_C = 100^\circ\text{C}$	I_{CEX}	- -	- -	10 35	mAdc
Emitter-Base Cutoff Current $V_{EB} = 15 \text{ V}$	I_{EBO}	-	-	10	mAdc
Emitter Floating Potential $V_{CB} = 15 \text{ V}, I_E = 0$	V_{EBF}	-	-	0.5	Vdc
Collector-Emitter Breakdown Voltage* $I_C = 500 \text{ mA}, I_B = 0$	BV_{CEO}	5	10	-	V
DC Current Transfer Ratio $I_C = 20 \text{ A}, V_{CE} = 2 \text{ V}$	h_{FE}	40	-	130	-
Collector-Emitter Saturation Voltage $I_C = 50 \text{ A}, I_B = 5 \text{ A}$	$V_{CE(sat)}$	-	.075	0.1	Vdc
Base-Emitter Voltage $I_C = 50 \text{ A}, I_B = 5 \text{ A}$	$V_{BE(sat)}$	-	0.85	1.0	Vdc
Common Emitter Cutoff Frequency $I_C = 20 \text{ A}, V_{CE} = 2 \text{ V}$	f_{ae}	3	4.5	-	kc
Rise Time $I_C = 20 \text{ A}, V_{CC} = 1.75 \text{ V}, I_{B(on)} = 2 \text{ A}$	t_r	-	18	25	μsec
Storage Time $V_{BE} = 6 \text{ V}, R_{be} = 10 \Omega$	t_s	-	15	20	μsec
Fall Time $V_{BE} = 6 \text{ V}, R_{be} = 10 \Omega$	t_f	-	10	15	μsec

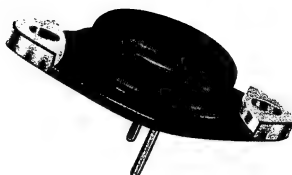
* To avoid excessive heating of the collector junction, perform these tests with an oscilloscope.

MOTOROLA HIGH-FREQUENCY POWER AMPLIFIER TRANSISTORS

EPITAXIAL PASSIVATED SILICON NPN

2N2947

2N2948



- High Pulse Breakdown Voltage :
2N2947 - 90V min 2N2948 - 80V min
- Power Output:
 $P_{out} = 15.0$ Watts at 50 mc
- Power Gain:
 $G_e = 10$ db typ at 50 mc with 15 Watts
RF Power Output

ABSOLUTE MAXIMUM RATINGS (Note 1)

Characteristic	Symbol	Rating		Unit
		2N2947	2N2948	
Collector-Base Voltage	V_{CBO}	60	40	Vdc
Collector-Emitter Voltage	V_{CES}	60	40	Vdc
Emitter - Base Voltage	V_{EB}	3	2	Vdc
Collector-Current (continuous)	I_C	1.5		Adc
Base-Current (continuous)	I_B	500		mAdc
Power Input (Nominal)	P_{in}	5.0		Watts
Power Output (Nominal)	P_{out}	20.0		Watts
Total Device Dissipation @ 25°C Case Temperature Derating Factor above 25°C	P_D	25.0		Watts
		167		mW/°C
Junction Temperature	T_J	175		°C
Storage Temperature	T_{stg}	-65 to + 175		°C

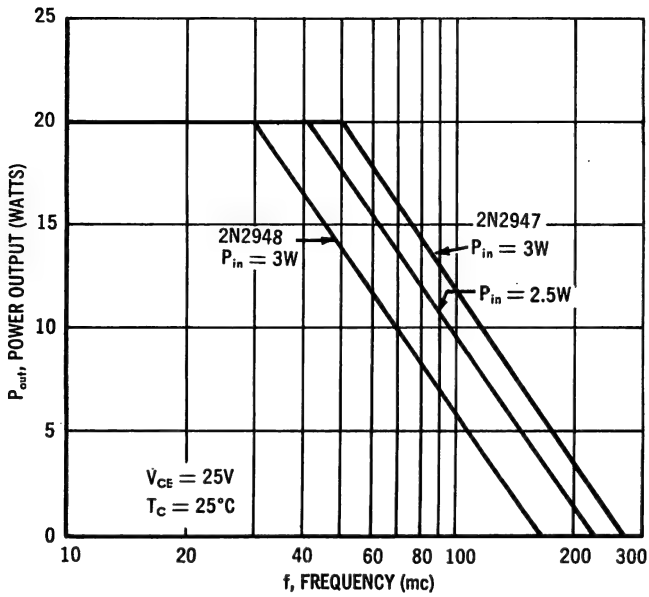
Note 1. The maximum ratings as given for DC conditions can be exceeded on a pulse basis. See electrical characteristics.

ELECTRICAL CHARACTERISTICS (T_A = 25°C unless otherwise noted)

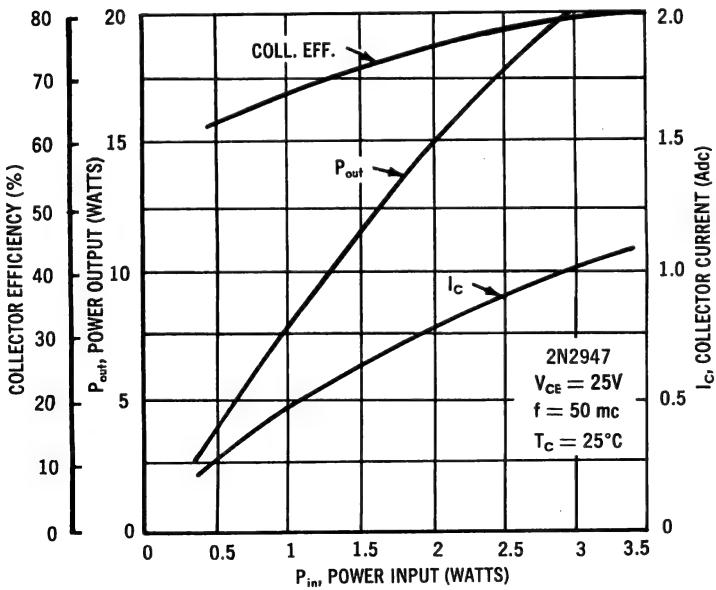
Characteristic	Symbol	Test Conditions	Min	Typ	Max	Unit
Collector-Emitter Sustain Voltage	V _{CES(sus)} *	2N2947: I _C = 0.250 A, R _{BE} = 0 2N2948: I _C = 0.250 A, R _{BE} = 0	90 80	120 100	--	Volts
Collector-Emitter-Open Base Sustain Voltage	V _{CEO(sus)} *	2N2947: I _C = 0.250 A, I _B = 0 2N2948: I _C = 0.250 A, I _B = 0	40 20	--	--	Volts
Collector-Emitter Current	I _{CES}	2N2947: V _{CE} = 60 Vdc, V _{BE} = 0 V _{CE} = 50 Vdc, V _{BE} = 0, T _C = 175°C 2N2948: V _{CE} = 40 Vdc, V _{BE} = 0 V _{CE} = 30 Vdc, V _{BE} = 0, T _C = 175°C	-- -- -- --	-- -- -- --	0.5 1.0 0.5 1.0	mAdc
Collector Cutoff Current	I _{CBO}	2N2947: V _{CB} = 50 Vdc, I _E = 0 2N2948: V _{CB} = 30 Vdc, I _E = 0	-- --	-- --	1 1	μAdc
Emitter Cutoff Current	I _{EBO}	2N2947: V _{EB} = 3 Vdc, I _C = 0 2N2948: V _{EB} = 2 Vdc, I _C = 0	-- --	-- --	100 100	μAdc
DC Current Gain	h _{FE}	2N2947: I _C = 400 mAdc, V _{CE} = 2Vdc 2N2948: I _C = 400 mAdc, V _{CE} = 2 Vdc Both Types: I _C = 1 Adc, V _{CE} = 2 Vdc	2.5 2.5 2.5	-- -- --	35 100 --	
Collector-Emitter Saturation Voltage	V _{CE(sat)}	I _C = 1.0 Adc, I _B = 500 mAdc	--	--	0.5	Vdc
Base-Emitter Saturation Voltage	V _{BE(sat)}	I _C = 1.0 Adc, I _B = 500 mAdc	--	--	2.0	Vdc
AC Current Gain	h _{fe}	V _{CE} = 2.0 Vdc, I _C = 400 mAdc, f = 50 mc	2.0	--	--	
Collector Output Capacitance	C _{ob}	V _{CB} = 25 Vdc, f _E = 0, f = 100 kc	--	--	60	pf
Power Input	P _{in}	2N2947 P _{out} = 15 W, f = 50 mc, V _{CE} = 25 Vdc	--	2.0	3.0	Watts
Efficiency	η	I _{C(max)} = 1 A	60	80	--	%
Power Input	P _{in}	2N2948 P _{out} = 15 W, f = 30 mc, V _{CE} = 25 Vdc	--	2.0	3.0	Watts
Efficiency	η	I _{C(max)} = 1.0 A	60	70	--	%

* Pulse Measurement: Pulse Width ≤ 100 μsec, Duty Cycle = 2%

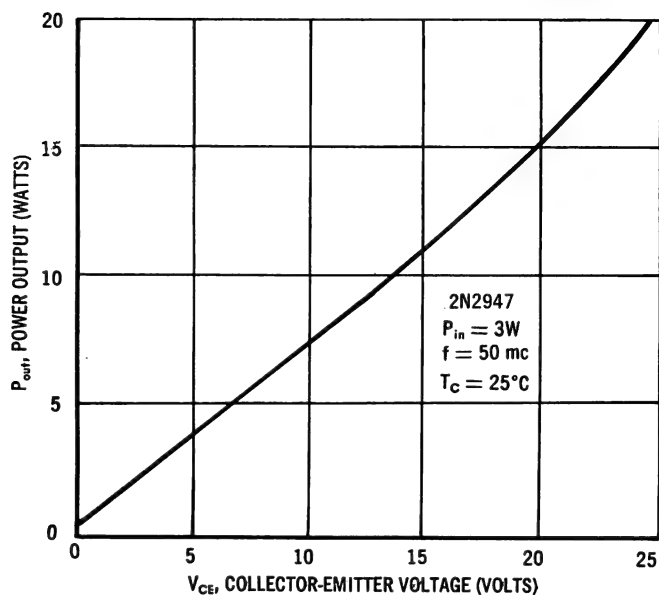
POWER OUTPUT versus FREQUENCY



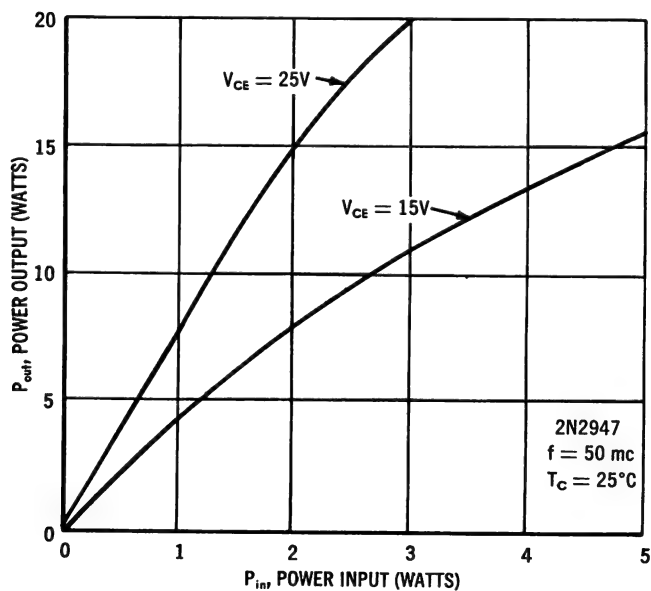
OUTPUT CHARACTERISTICS versus POWER INPUT



POWER OUTPUT versus COLLECTOR VOLTAGE



POWER OUTPUT versus POWER INPUT



MOTOROLA HIGH-FREQUENCY POWER AMPLIFIER TRANSISTORS

EPITAXIAL PASSIVATED SILICON NPN

2N2949



2N2950



- High Pulse Breakdown:

$$BV_{ces} = 85V$$

- Power Output:

$$P_{out} = 3.5 \text{ Watts at } 50 \text{ mc}$$

- Power Gain:

$$G_e = 12 \text{ db typical at } 50 \text{ mc with } 3.5 \text{ Watts of RF Power Output}$$

ABSOLUTE MAXIMUM RATINGS (Note 1)

Characteristic	Symbol	Rating		Unit
Collector-Base Voltage	V_{CBO}	60		Vdc
Collector-Emitter Voltage	V_{CES}	60		Vdc
Emitter - Base Voltage	V_{EB}	3.0		Vdc
Collector Current (Continuous)	I_C	0.7		Adc
Base Current (Continuous)	I_B	100		mAdc
RF Input Power (Nom)	P_{in}	1.0		Watt
RF Output Power (Nom)	P_{out}	5.0		Watts
Total Device Dissipation (25°C Case temperature) (Derating Factor above 25°C)	P_C	6.0 40		Watts mW/°C
Total Device Dissipation at 25° Ambient (Derating Factor above 25°C)	P_D	2N2949 0.5 3.33	2N2950 0.7 4.67	Watt mW/°C
Junction Temperature	T_J	175		°C
Storage Temperature	T_{stg}	-65 to +175		°C

NOTE 1:

The maximum ratings as given for DC conditions can be exceeded on a pulse basis. See Electrical Characteristics.

ELECTRICAL CHARACTERISTICS (T_A = 25°C unless otherwise noted)

Characteristic	Symbol	Test Conditions	Min	Typ	Max	Unit
Collector-Emitter Sustain Voltage	V _{CES(sus)} *	I _C =0.250 A, R _{BE} = 0	85	120	--	Volts
Collector Emitter-Open Base Sustain Voltage	V _{CEO(sus)} *	I _C =0.250A, I _B =0	40	--	--	Volts

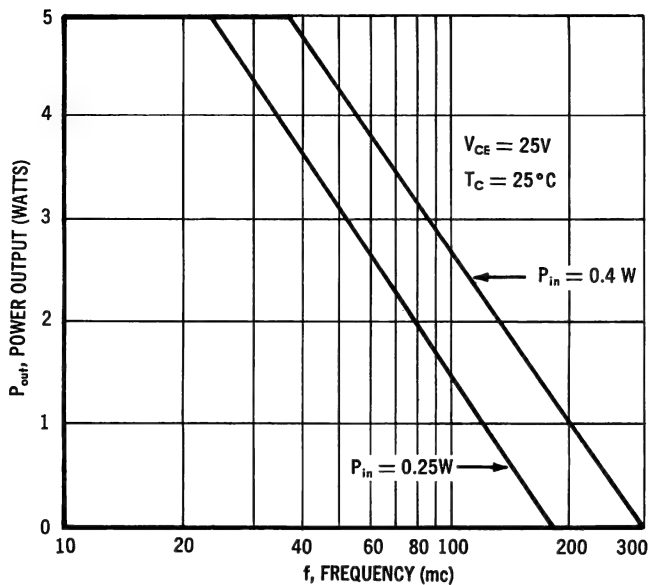
Collector-Emitter Current	I _{CES}	V _{CE} = 60 Vdc, V _{BE} =0	--	--	100	μAdc
		V _{CE} =50 Vdc, V _{BE} = 0 T _C = + 175°C	--	--	500	
Collector - Cutoff Current	I _{CBO}	V _{CB} =50 Vdc, I _E =0	--	--	0.1	μAdc
Emitter-Cutoff Current	I _{EBO}	V _{EB} = 3 Vdc, I _C =0	--	--	100	μAdc
DC Current Gain	h _{FE}	V _{CE} = 2.0 Vdc I _C = 40 mAdc	5.0	--	100	--
		V _{CE} =2.0 Vdc I _C = 400 mAdc	5.0	--	--	--
Collector - Emitter Saturation Voltage	V _{CE(sat)}	I _C = 400 mAdc, I _B = 80 mAdc	--	--	0.5	Vdc
Emitter-Base Saturation Voltage	V _{BE(sat)}	I _C =400 mAdc, I _B =80mAdc	--	--	2.0	Vdc

AC Current Gain	h _{fe}	V _{CE} =2.0 Vdc I _C =40 mAdc, f=50 mc	2.0	--	--	--
Collector Output Capacitance	C _{ob}	V _{CB} = 25 Vdc, I _E =0 f=100kc	--	--	20	pf

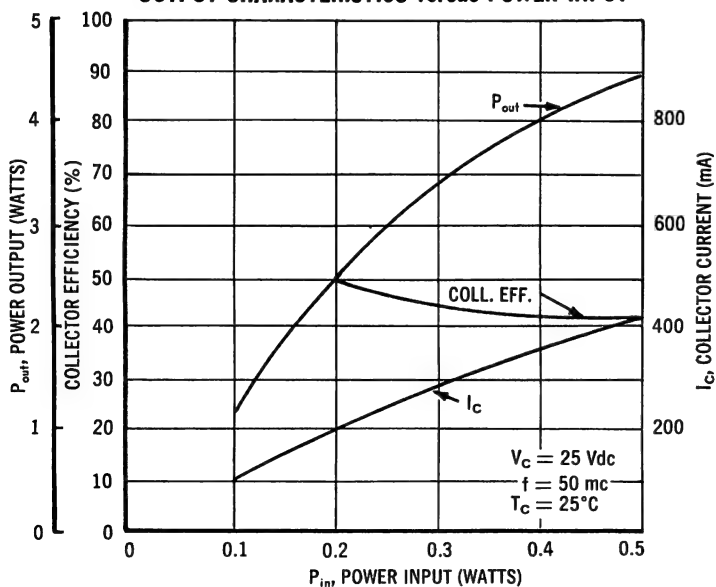
Power Input	P _{in}	P _{out} =3.5 watts, f=50 mc V _{CE} =25 Vdc, I _{C(max)} =325 mA,	--	--	0.35	Watt
Efficiency	η		43	--	--	%

* Pulse Width ≤ 100 μsec, Duty Cycle = 2%

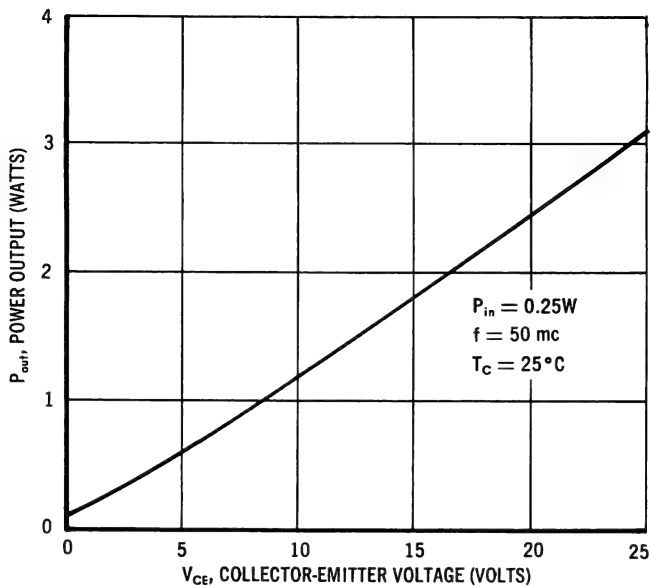
POWER OUTPUT versus FREQUENCY



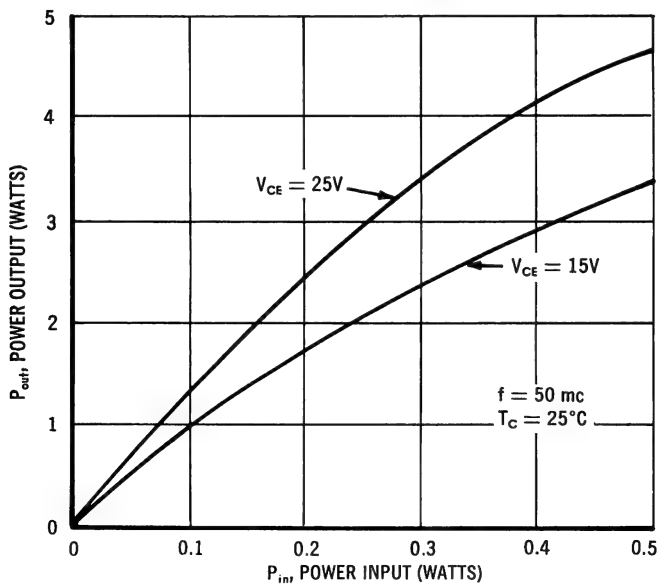
OUTPUT CHARACTERISTICS versus POWER INPUT



POWER OUTPUT versus COLLECTOR VOLTAGE



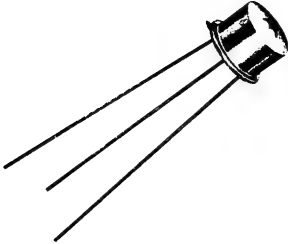
POWER OUTPUT versus POWER INPUT



MOTOROLA HIGH-FREQUENCY POWER AMPLIFIER TRANSISTORS

EPITAXIAL PASSIVATED SILICON NPN

2N2951



2N2952



- Power Output:
 $P_{out} = 600 \text{ mW at } 50 \text{ mc}$
- Power Gain:
 $G_e = 9 \text{ db typ at } 50 \text{ mc with } 600 \text{ mW of RF Power Output}$

ABSOLUTE MAXIMUM RATINGS (Note 1)

Characteristics	Symbol	Rating		Unit
Collector-Base Voltage	V_{CBO}	60		Vdc
Collector-Emitter Voltage	V_{CES}	60		Vdc
Emitter-Base Voltage	V_{EBO}	5		Vdc
Collector Current (continuous)	I_C	250		mAdc
Base Current (continuous)	I_B	50		mAdc
Total Device Dissipation (25°C Case Temperature) (Derate above 25°C)	P_C	2N2951	2N2952	Watts mW/°C
		3 20	1.8 12	
Total Device Dissipation (25°C Ambient Temperature) (Derate above 25°C)	P_D	0.8 5.33	0.5 3.33	mW/°C
Junction Temperature	T_J	-65 to 175		°C
Storage Temperature	T_{stg}	-65 to 175		°C

NOTE 1: The maximum ratings as given for D.C. conditions can be exceeded on a pulse basis. See Electrical Characteristics.

ELECTRICAL CHARACTERISTICS (At 25°C ambient unless otherwise noted)

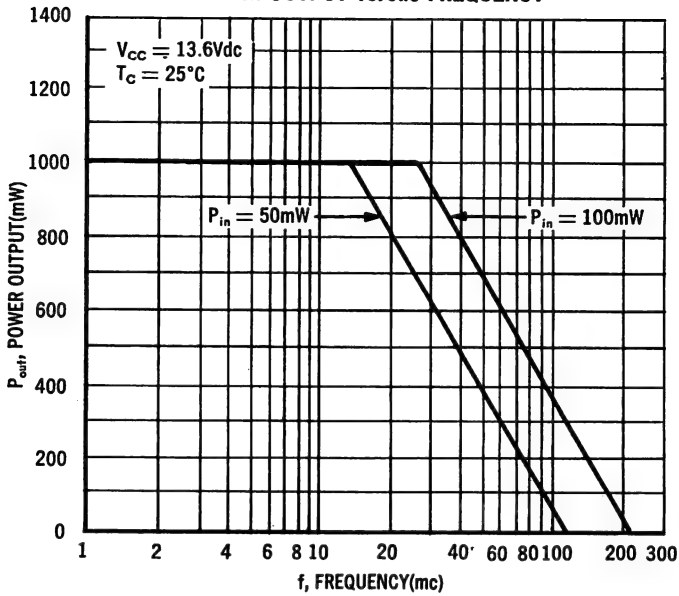
Characteristic	Symbol	Conditions	Min	Max	Unit
Collector-Emitter Current	I_{CES}	$V_{CE} = 60\text{Vdc}, V_{BE} = 0$	--	100	μAdc
		$V_{CE} = 50\text{Vdc}, V_{BE} = 0,$ $T_C = 175^\circ\text{C}$	--	500	μAdc
Collector Cutoff Current	I_{CBO}	$V_{GB} = 50\text{ Vdc}, I_E = 0$	--	0.1	μAdc
Emitter Cutoff Current	I_{EBO}	$V_{EB} = 5\text{ Vdc}, I_C = 0$	--	100	μAdc
DC Current Gain	h_{FE}	$I_C = 10\text{ mAdc}, V_{CE} = 10\text{ Vdc}$	20	150	--
		$I_C = 150\text{mAdc}, V_{CE} = 10\text{ Vdc}^*$	20	--	--
Collector-Emitter Saturation Voltage	$V_{CE(sat)}$	$I_C = 150\text{ mAdc}, I_B = 15\text{ mAdc}$	--	0.5	Vdc
Base-Emitter Saturation Voltage	$V_{BE(sat)}$	$I_C = 150\text{ mAdc}, I_B = 15\text{ mAdc}$	--	2.0	Vdc

Collector-Emitter Sustain Voltage	$V_{CES(sus)}^*$	$I_C = 100\text{ mA}, R_{BE} = 0$	30	--	Volts
Collector-Emitter Open Base Sustain Voltage	$V_{CEO(sus)}^*$	$I_C = 100\text{ mA}, I_B = 0$	20	--	Volts
AC Current Gain	$ h_{fe} $	$V_{CE} = 10\text{ Vdc}, I_C = 10\text{ mAdc}$ $f = 50\text{ mc}$	4.0	--	--
Collector Output Capacitance	C_{ob}	$V_{CB} = 10\text{ Vdc}, I_E = 0, f = 100\text{kc}$	--	8	pf

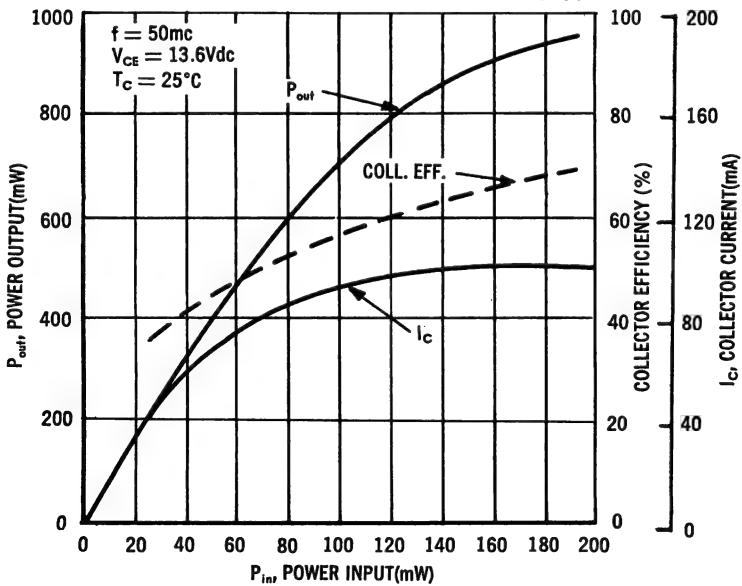
Power Input	P_{in}	$P_{out} = 600\text{ mW}$ $f = 50\text{ mc}$ $V_{CE} = 13.6\text{ Vdc}$ $I_{C(max)} = 125\text{ mA}$	--	100	mW
Efficiency	η		35	--	%

*Pulse $\leq 100\text{ nsec}$, Duty Cycle = 2%

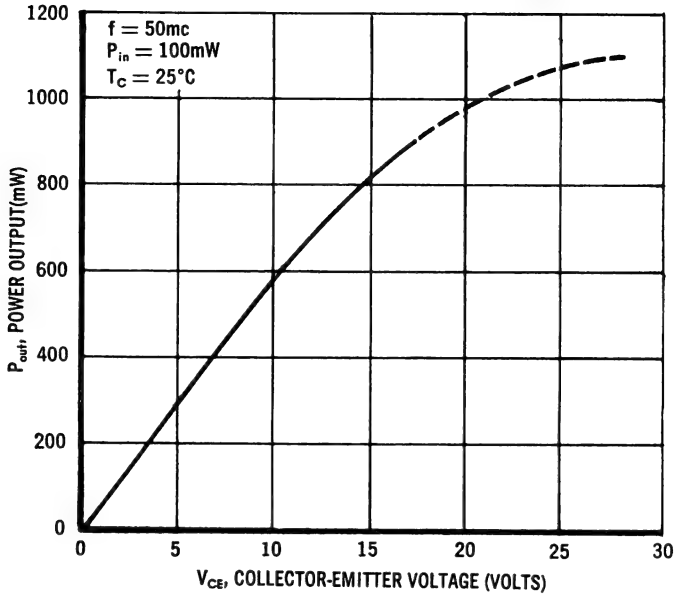
POWER OUTPUT versus FREQUENCY



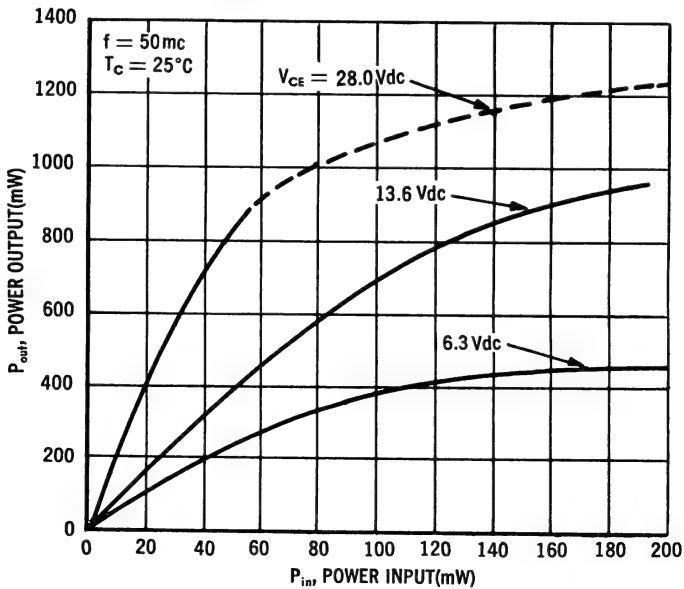
OUTPUT CHARACTERISTICS versus POWER INPUT



POWER OUTPUT versus COLLECTOR VOLTAGE



POWER OUTPUT versus POWER INPUT



NOTES

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